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News
Previewing the Wireless Systems Design Conference

Design Feature
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Product Technology
Top Products of 2003

Load-Pull Tuners Are Frequency Selective

Test & Measurement Issue

WORLDWIDE
CONNECTIONS

Amplifiers for Every Application

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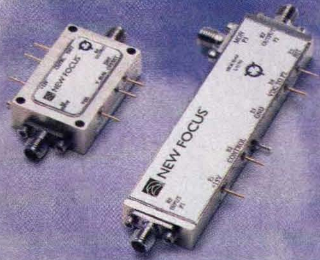

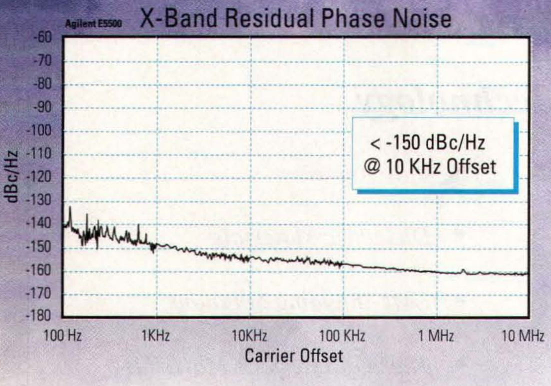

Broadband Power Amplifiers NEW

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ
JCA018-3000	2.0–18.0	25	6.0	2.0	23	28
JCA218-3001	2.0–18.0	25	6.0	2.0	25	30
JCA218-3002	2.0–18.0	25	6.0	2.0	27	32
JCA218-4000	2.0–18.0	30	6.0	2.0	23	28
JCA218-4001	2.0–18.0	30	6.0	2.0	25	30
JCA218-4002	2.0–18.0	30	6.0	2.0	27	32
JCA218-5000	2.0–18.0	35	6.0	2.0	23	28
JCA218-5001	2.0–18.0	35	6.0	2.0	25	30
JCA218-5002	2.0–18.0	35	6.0	2.0	27	32

Power Amplifiers

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ
JCA12-P01	1.35–1.85	35	4.0	1.0	33	41
JCA34-P02	3.1–3.5	40	4.5	1.0	37	45
JCA56-P01	5.9–6.4	30	5.0	1.0	34	42
JCA812-P03	8.0–12.0	40	5.0	1.5	33	40
JCA1218-P02	12.0–18.0	22	4.0	2.0	25	35

Low Noise Amplifiers

Model	Freq. Range GHz	Gain dB min	N/F dB max	Flatness +/-dB	1 dB Comp. pt. dBm min	3rd Order ICP typ
JCA12-1000	1.2–1.6	25	0.8	0.5	10	20
JCA12-3001	1.0–2.0	40	0.8	1.0	10	20
JCA23-302	2.2–2.3	30	0.8	0.5	10	20
JCA34-301	3.7–4.2	30	1.0	0.5	10	20
JCA78-300	7.25–7.75	27	1.2	0.5	13	23
JCA910-3000	9.0–9.5	25	1.3	0.5	13	23
JCA1112-3000	11.7–12.2	27	1.4	0.5	13	23
JCA1415-3001	14.4–15.4	35	1.6	1.0	14	24
JCA1819-3001	18.1–18.6	25	2.0	0.5	10	20
JCA2021-3001	20.2–21.2	25	2.5	0.5	10	20

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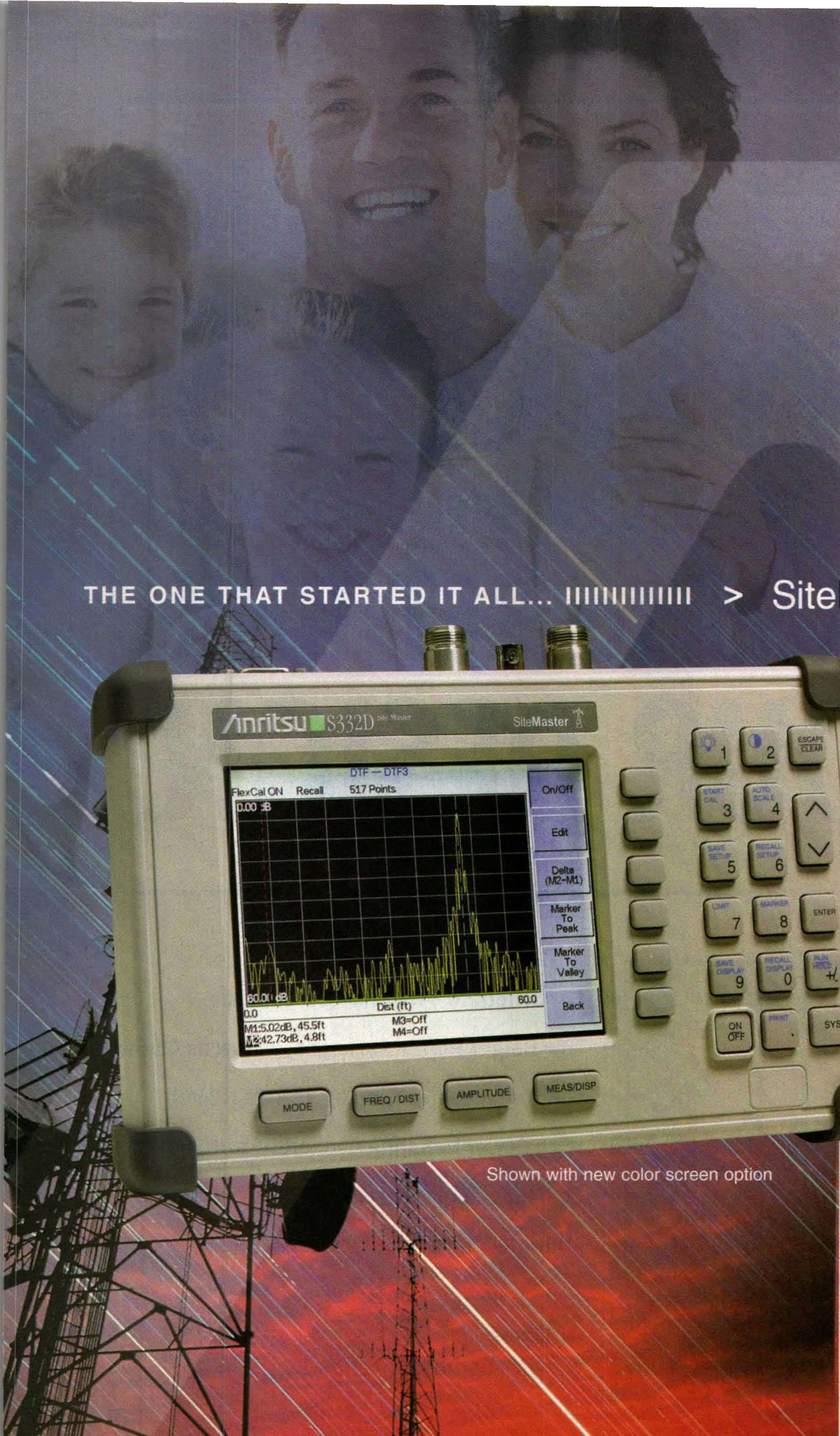

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Omission Correction

▶▶IN OUR ARTICLE "Enhance The Design Of LTCC RF Modules" (September 2003, p. 90), we regret our failure to cite the original reference for the design procedure used to devise an equivalent circuit for a lumped element directional coupler. The original work was provided in:

Taek-Young Song, Jae-Ho Kim, Sang-Hyuk Kim, Jae-Bong Lim, and Jun-Seok Park, "Design of a Novel Lumped Element Backward Directional Coupler Based on Parallel Coupled-Line Theory," *IEEE International Microwave Symposium Digest*, pp. 213-216, Seattle, WA, June 2002.

The authors apologize for their failure and wish to convey that the omission was purely an editing oversight.

Lawrence Williams
Sean Kim
Ansoft Corp.

October Editorial

▶▶RE YOUR EDITORIAL, "DDS Gains Wider Application" (October 2003, p. 17): Unless NASA did something foolish, I believe that the concept of DDS has been in use since at least 1979 from a design I did and for which NASA eventually had a contract awarded at Martin Marietta in Denver.

At that period in which I designed, and should have had in place, the first JTAG uWave system in the world, the DDS system included DDM (Direct Digital Modulation) for nPSKm, nFM-mod, and AM under digital control (also, of course, the other multiplicity of CW modes).

My opinion is that your opinion is at least 25 years out of date. Please advise me if I am in error. It seems that all of our spacecraft failures are unrelated.

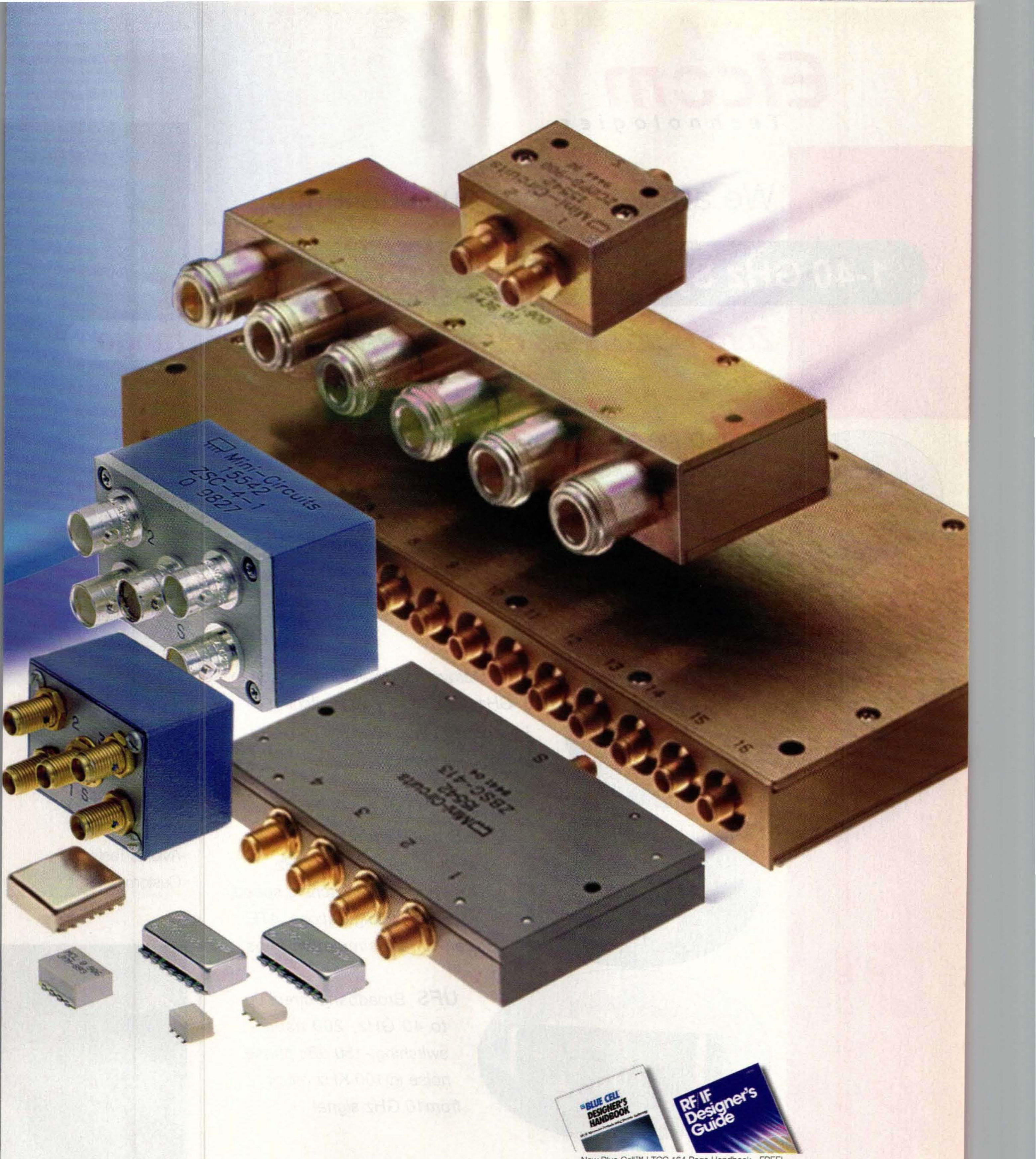
Ron Reasoner

Editor's Note

▶▶IN THE ARTICLE "Make Accurate Pulsed S-Parameter Measurements" by Loren Betts, which appeared on p. 72 of the November 2003 issue of *Microwaves & RF*, the same figure was mistakenly published as Figures 11, 12, and 13. They should have been different figures. The correct figures have been posted online in the Internet version of the article, which readers can see on our newly revamped website at www.planetee.com. In addition, the new website also has numerous features which we are sure will be of interest to the readers of the magazine.

We apologize to the author, Loren Betts, and to our readers for the oversight. The figures are quite similar, so the error was an easy one to make.

The Editors of Microwaves & RF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Facing The Next Test Challenge

TEST-EQUIPMENT DEVELOPERS are to be applauded for their knack of anticipating the industry's high-frequency measurement needs. In spite of the plethora of wireless-communications formats and standards, including the various flavors of cellular radios, Bluetooth, and the many configurations of wireless local-area networks (WLANs) to name a few, suppliers of test solutions have either been ready with targeted instrumentation or, in the worst case, responded quickly to an emerging measurement need.

UWB represents a challenge that the test-equipment development community has not faced before.

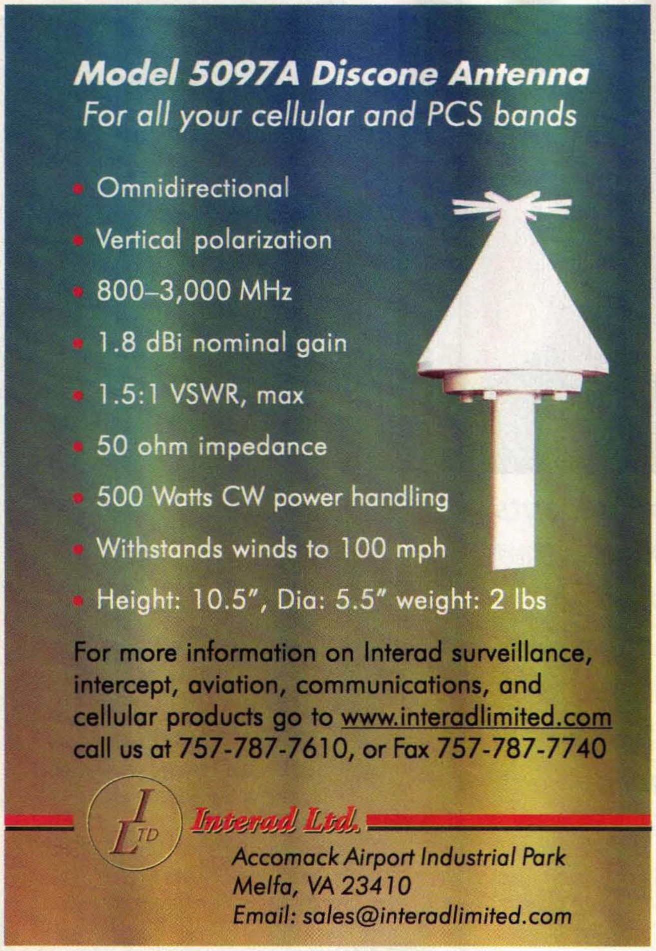
Even when wireless standards have changed directions, as in the case with WLANs, test engineers have been able to follow the most erratic product-development paths—often by participating in standards-development committees—and be ready with practical solutions when the market demanded them. Sometimes these solutions have come in the form of dedicated test equipment; sometimes they have been made available as modules or software "personality" upgrades for existing equipment.

When searching for test equipment for one of the newer emerging wireless-communications formats, however, the solutions are more approximations than exact fits. That emerging technology, ultrawideband (UWB), certainly represents a challenge that the test-equipment development community has not faced before and is actually closer to requirements typically established by the surveillance community. Two years ago, the US Federal Communications Commission (FCC) gave its preliminary blessing to the use of UWB technology (by means of its First Report and Order issued on February 14, 2002), with the promise to monitor its potential interference effects on existing, traditional users of RF signals, such as cellular telephones, Global Positioning Systems (GPS), and WLANs. The FCC (www.fcc.org) actually established three different groups of UWB-based products: imaging systems (such as ground-penetrating radars), vehicle radar systems, and communications and measurement systems. This last group, which is of greatest concern to existing standards and unlicensed users of RF energy, has been restricted to the frequency band from 3.1 to 10.6 GHz.

Although debate continues, with or without the IEEE, on just how to implement an industry-wide standard for unlicensed UWB communications, test-equipment developers are currently stymied as to how to evaluate UWB gear. Most test tools are designed for relatively narrow channels, even as much as 20 MHz in wideband CDMA systems. But capturing several GHz of bandwidth in one shot represents a whole new problem. Still, test-equipment developers have never ceased to amaze when meeting new challenges, even one as daunting as this.

Jack Browne

Publisher/Editor



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AFS2-00700080-06-10P-6	.7–.8	28	0.50	0.60	1.5:1	1.5:1	+10	90
AFS2-00800100-05-10P-6	.8–1	30	0.50	0.50	1.5:1	1.5:1	+10	90
AFS3-01200160-05-13P-6	1.2–1.6	40	0.50	0.50	1.5:1	1.5:1	+13	150
AFS3-01400170-06-13P-6	1.4–1.7	40	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-01500180-06-13P-6	1.5–1.8	40	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-01500250-06-13P-6	1.5–2.5	38	1.00	0.60	1.8:1	1.8:1	+13	150
AFS3-01700190-06-13P-6	1.7–1.9	38	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-01800220-06-13P-6	1.8–2.2	38	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-02200230-06-13P-4	2.2–2.3	38	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-02300270-06-13P-6	2.3–2.7	36	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-02700290-06-13P-6	2.7–2.9	32	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-02900310-06-13P-6	2.9–3.1	32	0.50	0.60	1.5:1	1.5:1	+13	150
AFS3-03100350-06-10P-4	3.1–3.5	29	0.50	0.60	1.5:1	1.5:1	+10	150
AFS4-03400420-10-13P-6	3.4–4.2	40	0.50	1.00	1.5:1	1.5:1	+13	200
AFS3-04400510-07-S-4	4.4–5.1	30	0.50	0.70	1.5:1	1.5:1	+10	100
AFS3-04500480-07-S-4	4.5–4.8	30	0.50	0.70	1.5:1	1.5:1	+10	100
AFS3-05200600-07-10P-4	5.2–6	30	0.50	0.70	1.5:1	1.5:1	+10	100
AFS3-05400590-07-S-4	5.4–5.9	30	0.50	0.70	1.5:1	1.5:1	+10	100
AFS3-05800670-07-S-4	5.8–6.7	30	0.50	0.70	1.5:1	1.5:1	+10	100
AFS3-07250775-06-10P-4	7.25–7.75	30	0.50	0.60	1.5:1	1.5:1	+10	100
AFS3-07900840-07-S-4	7.9–8.4	30	0.50	0.70	1.5:1	1.5:1	+10	100
AFS4-08500960-08-S-4	8.5–9.6	32	0.75	0.80	1.5:1	1.5:1	+10	125
AFS3-09001100-09-S-4	9–11	26	0.50	0.90	1.5:1	1.5:1	+10	100
AFS4-09001100-09-S-4	9–11	32	0.75	0.90	1.5:1	1.5:1	+10	125
AFS4-10951175-09-S-4	10.95–11.75	32	0.75	0.90	1.5:1	1.5:1	+10	125
AFS4-11701220-09-5P-4	11.7–12.2	32	0.75	0.90	1.5:1	1.5:1	+10	125
AFS2-12201280-14-5P-2	12.2–12.8	14	0.75	1.40	1.4:1	1.5:1	+5	80
AFS4-12201280-13-12P-4	12.2–12.8	25	1.50	1.30	2.0:1	2.0:1	+12	200
AFS4-12701330-15-10P-4	12.7–13.3	30	0.75	1.50	1.5:1	1.5:1	+10	175
AFS4-13201400-16-10P-4	13.2–14	30	0.75	1.60	1.5:1	1.5:1	+10	175
AFS4-14001450-15-10P-4	14–14.5	30	0.75	1.50	1.5:1	1.5:1	+10	175
AFS4-20202120-25-8P-4	20.2–21.2	24	1.00	2.50	1.5:1	1.5:1	+8	175
AFS4-21202400-28-10P-4	21.2–24	23	1.00	2.80	2.0:1	2.0:1	+10	100
OCTAVE BAND AMPLIFIERS								
AFS3-00120025-09-10P-4	.12–.25	38	0.50	0.9	2.0:1	2.0:1	+10	125
AFS3-00250050-08-10P-4	.25–.5	38	0.50	0.8	2.0:1	2.0:1	+10	125
AFS3-00500100-06-10P-6	.5–1	38	0.75	0.6	2.0:1	1.5:1	+10	150
AFS3-01000200-05-10P-6	1–2	38	1.00	0.5	2.0:1	2.0:1	+10	150
AFS3-01200240-06-10P-6	1.2–2.4	34	1.00	0.6	2.0:1	2.0:1	+10	150
AFS3-02000400-06-10P-4	2–4	32	1.00	0.6	2.0:1	2.0:1	+10	125
AFS3-02600520-10-10P-4	2.6–5.2	28	1.00	1.0	2.0:1	2.0:1	+10	125
AFS3-04000800-07-10P-4	4–8	28	1.00	0.7	2.0:1	2.0:1	+10	125
AFS3-08001200-09-10P-4	8–12	26	1.00	0.9	2.0:1	2.0:1	+10	125
AFS3-08001600-15-8P-4	8–16	28	1.00	1.5	2.0:1	2.0:1	+8	100
AFS4-12002400-30-10P-4	12–24	24	2.00	3.0	2.0:1	2.0:1	+10	85
AFS4-12001800-18-10P-4	12–18	28	1.50	1.8	2.0:1	2.0:1	+10	125
AFS3-18002650-30-8P-4	18–26.5	18	1.75	3.0	2.2:1	2.2:1	+8	125
MULTIOCTAVE BAND AMPLIFIERS								
AFS1-00040200-12-10P-4	.04–2	15	1.50	1.2	2.5:1	2.0:1	+10	75
AFS3-00300140-09-10P-4	.3–1.4	38	1.00	0.9	2.0:1	2.0:1	+10	125
AFS2-00400350-12-10P-4	.4–3.5	22	1.50	1.2	2.0:1	2.0:1	+10	80
AFS3-00500200-08-15P-4	.5–2	38	1.00	0.8	2.0:1	2.0:1	+15	125
AFS3-01000400-10-10P-4	1–4	30	1.50	1.0	2.0:1	2.0:1	+10	125
AFS3-02000800-09-10P-4	2–8	26	1.00	1.0	2.0:1	2.0:1	+10	125
AFS4-02001800-23-10P-4	2–18	25	2.00	2.3	2.0:1	2.0:1	+10	175
AFS4-06001800-22-10P-4	6–18	25	2.00	2.2	2.0:1	2.0:1	+10	125
AFS4-08001800-22-10P-4	8–18	28	2.00	2.2	2.0:1	2.0:1	+10	125
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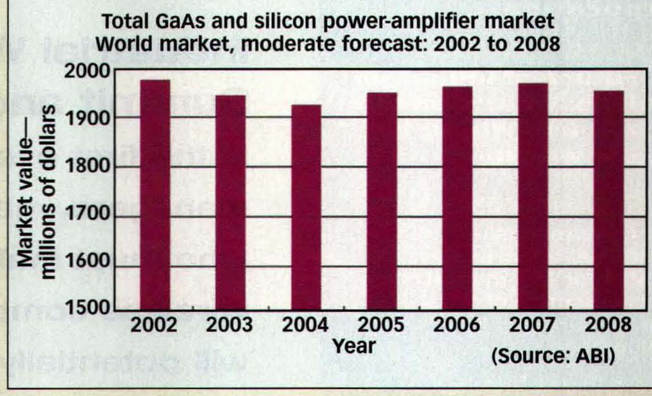
AFS3-00100100-09-10P-4	.1–1	38	1.00	0.9	2.0:1	2.0:1	+10	125
AFS3-00100200-10-15P-4	.1–2	38	1.00	1.0	2.0:1	2.0:1	+15	150
AFS3-00100300-12-10P-4	.1–3	32	1.00	1.2	2.0:1	2.0:1	+10	125
AFS3-00100400-13-10P-4	.1–4	28	1.00	1.3	2.0:1	2.0:1	+10	125
AFS3-00100600-13-10P-4	.1–6	28	1.25	1.3	2.0:1	2.0:1	+10	125
AFS3-00100800-14-10P-4	.1–8	28	1.50	1.4	2.0:1	2.0:1	+10	125
AFS4-00101200-22-10P-4	.1–12	30	1.50	2.2	2.0:1	2.0:1	+10	150
AFS4-00101400-23-10P-4	.1–14	24	2.00	2.3	2.5:1	2.5:1	+10	200
AFS4-00101800-25-S-4	.1–18	25	2.00	2.5	2.5:1	2.5:1	+10	175
AFS4-00102000-30-10P-4	.1–20	20	2.50	3.0	2.5:1	2.5:1	+10	125
AFS4-00102650-42-8P-4	.1–26.5	22	2.50	4.2	2.5:1	2.5:1	+8	135

Note: Noise figure increases below 500 MHz in bands greater than 0.1-10 GHz.

The Power-Semiconductor Market Is Expected To Remain Flat

OYSTER BAY, NY—The overall market for gallium-arsenide (GaAs) and silicon-based power semiconductors will remain relatively flat over the next five years, hovering at about the $2 billion mark, according to recent findings from technology market research firm ABI **(see figure)**. In the firm's aggressive forecast, the market will rise slightly at 2 percent, compounded annually, over the next five years. With this market still heavily dependent on the ailing cellular industry, robust growth is limited to niche markets.



For year-end 2003, ABI anticipates that cellular applications will comprise 64 percent of the overall market for GaAs and silicon power semiconductors. Currently, this is split almost evenly between handsets and infrastructure. Over time, however, the cellular industry will gradually represent a diminishing share of the overall market.

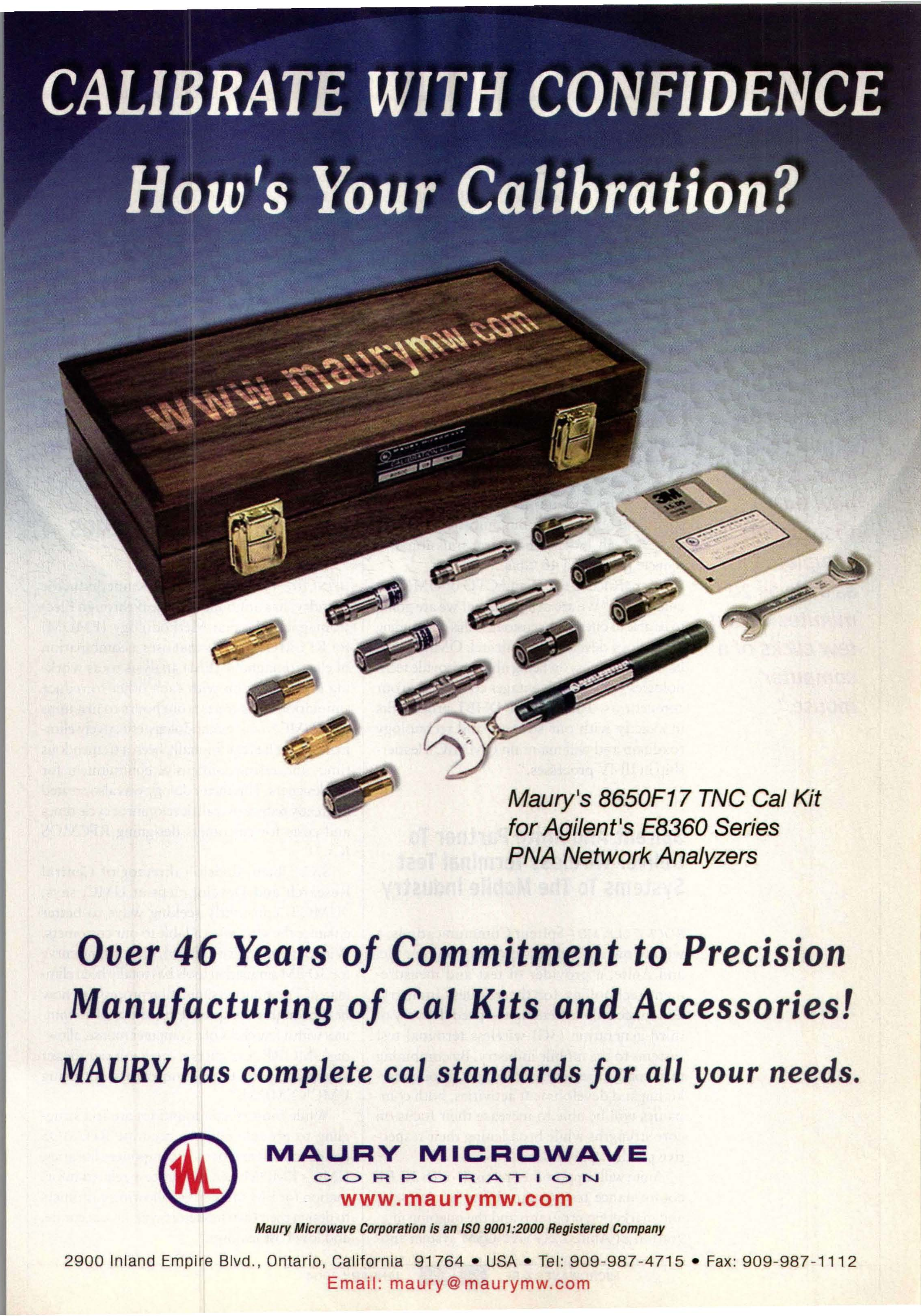
"By 2008, cellular handsets and infrastructure will represent about 50 percent of the broader market for RF power devices," explains Edward Rerisi, ABI's director of research. "Niche applications—including Wi-Fi, satellite, and military applications—will help buoy the overall market as the cellular segment decreases. However, their combined impact will not boost the market sufficiently enough to reestablish robust industry growth."

In 2008, ABI anticipates 55 percent of the market to be GaAs and 45 percent of the market to be silicon-based.

WLAN Hardware Market Was Up 11 Percent To $658M In Q3

LONDON, ENGLAND—Worldwide wireless-local-area-network (WLAN) hardware revenue topped $658 million in the third quarter. This is up 11 percent from the second quarter, and is projected to grow to $3.6 billion in 2006, more than doubling from $1.6 billion in 2002, according to Infonetics Research's quarterly worldwide market-share and forecast service, Wireless LAN Hardware. Growth in the market is being driven by the widespread popularity of WiFi across enterprise, consumer, and hotspot markets as mobility becomes ingrained as both a way of working and in consumer lifestyles.

"One of the most interesting aspects of this market is the wireless LAN switch segment, which is immature but has strong potential. Port shipments grew 95 percent to 23,000 this quarter, and revenue grew 100 percent to $12 million," comments Infonetics Research's Richard Webb, lead analyst of the report. "This hearty growth is due mostly to the fact that this quarter marks the first quarter of recognized shipments and revenue for several vendors. But we forecast healthy quarterly revenue and port growth in the double-digit percentages through 2004, and annual revenue and port growth in the double- to triple-digit percentages through 2006, at which time worldwide revenue for wireless LAN switches will reach $169 million. However, this is a conservative forecast based on initial results. If the wireless LAN switch vendors can get into the larger enterprise, this figure will be much larger.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Alcatel And OMMIC Cooperate In Microelectronic Project

PARIS, FRANCE—Alcatel and OMMIC announced that they are collaborating on the transfer of Alcatel's advanced indium-phosphide (InP) heterojunction-bipolar-transistor (HBT) technology to OMMIC's industrial clean rooms at its center in Limeil-Brévannes, France. This transfer will complete OMMIC's commercial portfolio of III-V technologies and will provide Alcatel's System designers with a qualified source to develop 40 Gb/s transmission systems for core networks.

InP, now well established as the choice material for long wavelength (1.3 to 1.6 μm) optoelectronic devices, is receiving an increasing interest for its potential in microelectronic applications, ranging from millimeter-wave space and terrestrial communication systems, to very high bit-rate fiber transmission systems (e.g., 10 and 40 Gb/s).

Marc Rocchi, COO and CTO of OMMIC, comments, "We are delighted that we are going to be able to offer to our customers this outstanding technology developed by Alcatel. OMMIC has based its business on being able to provide technologies with real advantages compared to our competitors. This new InP DHBT process fits in exactly with our strategy and technology roadmap and will maintain OMMIC's leadership in III-IV processes."

"The inductor simulation process can now be thoroughly completed in as little as 20 minutes with a few clicks of a computer mouse."

Spirent And Anite Partner To Deliver Wireless Terminal Test Systems To The Mobile Industry

ROCKVILLE, MD—Spirent Communications, a wholly owned business group of Spirent plc, and Anite, a provider of test and measurement technology for the wireless industry, announced a partnership to speed delivery of third-generation (3G) wireless terminal test systems to the mobile industry. By combining their complementary technologies in joint marketing and development activities, both companies will be able to increase their focus on core strengths while broadening their respective product portfolios.

Anite will support the Spirent U-ATS 3G RF conformance test system through joint sales and marketing programs and the ongoing integration of Anite's 2G SAT GSM system simulator into the Spirent system, to enable 3GPP Radio Resource Management (RRM) test case coverage.

Spirent will support the Anite U-SAT 3G protocol conformance test system through joint sales and marketing programs and the ongoing development of the CS100 platform. The agreement will also allow Anite to increase its development focus on the product, which already provides broad coverage of the signalling test cases in the 3GPP terminal conformance specifications.

In addition, under the terms of the partnership, Anite will cease production of its own U-RAMS 3G RF conformance tester, and Spirent will not produce its own protocol conformance tester.

UMC Introduces New RFCMOS Design Methodology

HSINCHU, TAIWAN—UMC, a semiconductor foundry, has announced a breakthrough Electromagnetic Design Methodology (EMDM) for RFCMOS designs that uses a combination of electromagnetic (EM) analysis tools working in conjunction with each other to reduce simulation cycle times from hours to just minutes. UMC's new methodology effectively eliminates what has traditionally been a tremendous time- and resource-intensive commitment for RF designers. This methodology was also created to greatly reduce overall development cycle times and costs for customers designing RFCMOS ICs.

S.C. Chien, division director of Central Research and Development at UMC, says, "UMC is constantly seeking ways to better enhance the services available to our customers. With our innovative EMDM, the learning curve for 3D EM simulation tools has totally been eliminated. The inductor simulation process can now be thoroughly completed in as little as 20 minutes with a few clicks of a computer mouse, allowing UMC RF customers to realize a significant advantage over their competitors not using UMC's EMDM."

While most silicon foundries are still struggling to provide reliable, accurate RFCMOS design models and passive component libraries, UMC's EMDM includes process-related information for EM simulation, allowing customers to design their own inductors with fast, accurate, and low-cost features.

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S4W2	S4W5	N4W5	4	±0.40
S5W2	S5W5	N5W5	5	±0.40
S6W2	S6W5	N6W5	6	±0.40
S7W2	S7W5	N7W5	7	±0.60
S8W2	S8W5	N8W5	8	±0.60
S9W2	S9W5	N9W5	9	±0.60
S10W2	S10W5	N10W5	10	±0.60
S12W2	S12W5	N12W5	12	±0.60
S15W2	S15W5	N15W5	15	±0.60
S20W2	S20W5	N20W5	20	±0.60
S30W2	S30W5	N30W5	30	±0.85
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The Success Of CDMA2000 Technology Was Examined At The 2003 CDMA Americas Congress

COSTA MESA, CA—More than 1000 wireless professionals from the pan-American region attended the eighth annual CDMA Americas Congress in Miami, FL to learn about the latest developments of CDMA2000®. Organized by the CDMA Development Group (CDG) in conjunction with the Institute for International Research (IIR), the conference stated the case for continuing the advancement of CDMA2000 as a wireless solution to provide increased speed and flexibility for voice and data services within the Americas. Participants included operators, vendors, and content providers from around the world, with a significant increase in Latin American attendance, demonstrating growing interest in CDMA2000 in the region.

"CDMA2000 offers the most advanced services today."

"CDMA2000 is a flexible technology, allowing carriers to offer competitive solutions for affordable voice services and high-speed access to data applications," says Perry LaForge, executive director for the CDG. "Consumers and enterprise users are being introduced to a broad range of applications and services, while carriers are finding CDMA2000 to be a source of new revenue streams, strengthening their capacity and allowing them to differentiate their products and services. With more than 75 million 3G wireless subscribers on 80 networks throughout the world, CDMA2000 offers the most advanced services today."

Highlights from the event included up-to-the-minute technical information on CDMA2000 technology evolution, an extensive exhibition, and the latest in next-generation devices. The 2003 CDMA Americas Congress presented strategies for realizing revenue, recent Latin America success stories, interactive workshops, and the ninth annual CDMA Test Forum. The CDG also held a press conference where operators discussed their experiences and future plans with CDMA2000.

Kudos

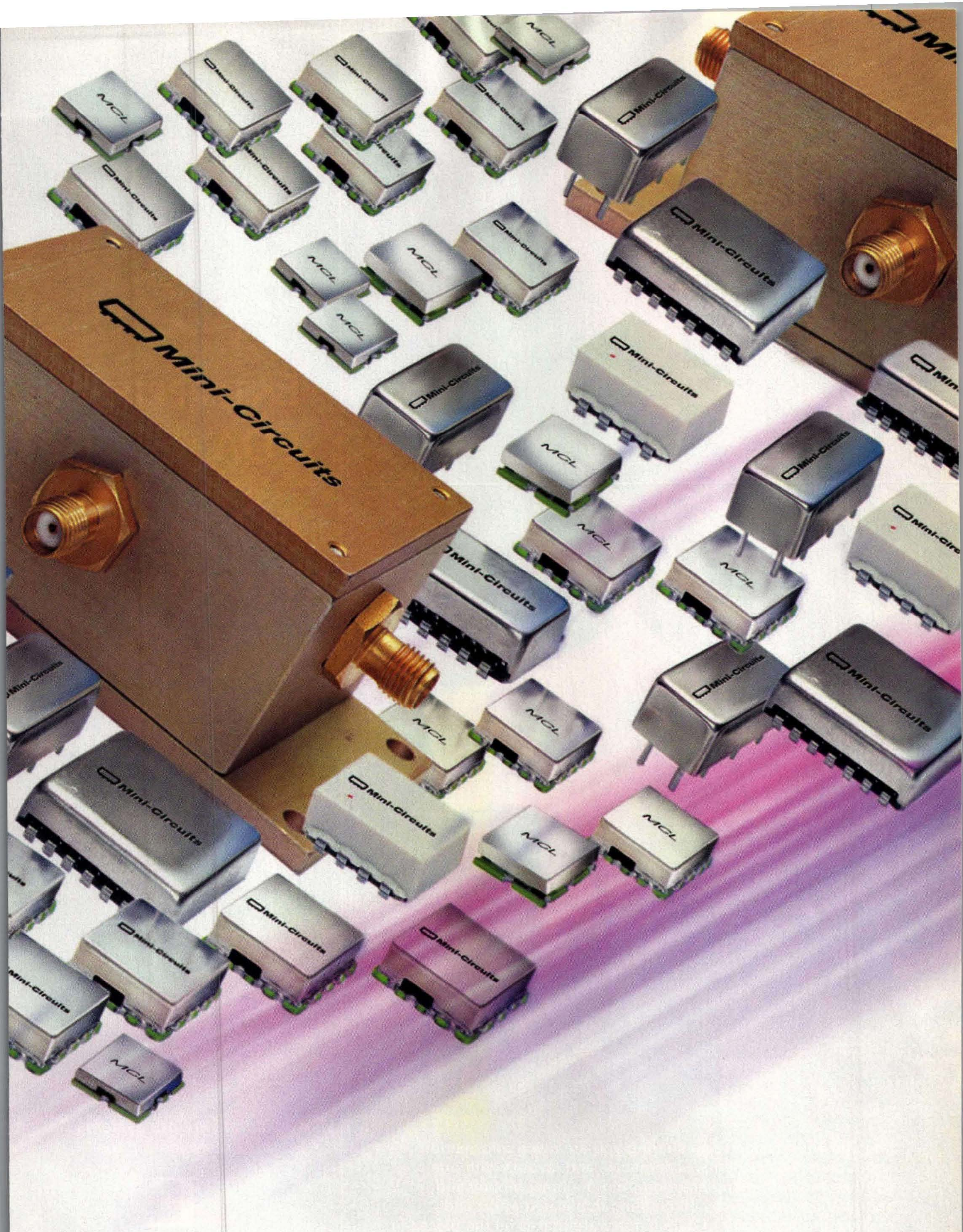
HOUSTON, TX—Mimix Broadband, Inc., a fabless semiconductor company that supplies monolithic microwave integrated circuits (MMICs), has announced that they have been recommended for registration to ISO 9001:2000, the internationally recognized quality management system standard administered by the International Organization for Standardization. Det Norske Veritas (DNV) Certification, Inc. has recommended registration for Mimix to ISO 9001:2000. Mimix was previously registered by DNV to ISO 9001:1994.

WALLINGFORD, CT—Times Microwave Systems was recently recognized by Part-15.org (www.part-15.org) as their "Manufacturer of the Year," during Part-15.org's recent conference in Dallas, TX. This award is given to a manufacturer for exceptional technical support, innovation of new products and dedication and involvement in education, training, and lecture programs.

WARREN, NJ—ANADIGICS, Inc., a supplier of wireless and broadband solutions, announced that Ronald Rosenzweig, chairman of the board and co-founder of ANADIGICS, was among six select industry pioneers honored at the 9th Annual Compound Semi Industry Outlook Conference held in Dallas, TX. The award presentation took place on Tuesday, December 16th during the event's "Pioneer Awards" gala at the Westin Galleria Hotel. Mr. Rosenzweig also addressed attendees as a featured panelist during the event, which attracted senior managers of advanced semiconductor companies and financial analysts covering the communications sector.

RICHARDSON, TX—TriQuint Semiconductor's vice president, David McQuiddy, was honored at the 9th Annual Compound Semi Industry Outlook Conference held on December 16 in Dallas, TX. McQuiddy was feted for his life-long contributions to the field, and received one of six Industry Pioneer awards.

MEDWAY, MA—NARTE (the National Association of Radio Telecommunications Engineers, Inc.) presented Marconi-Bell Awards to Charles Glass of the Commerce Department and Scott Harris of the law firm Harris Wiltshire & Grannis. Glass and Harris were honored for their work in helping to negotiate an international government/industry agreement on 5-GHz WiFi that was approved at the World Radiocommunication Conference in Geneva, Switzerland in 2003. Glass represented the public sector and Harris represented the private sector in reaching a solution based on "dynamic frequency selection," or DFS, which enables spectrum to be shared by industry and government while keeping sensitive Defense Department systems protected from interference. MRF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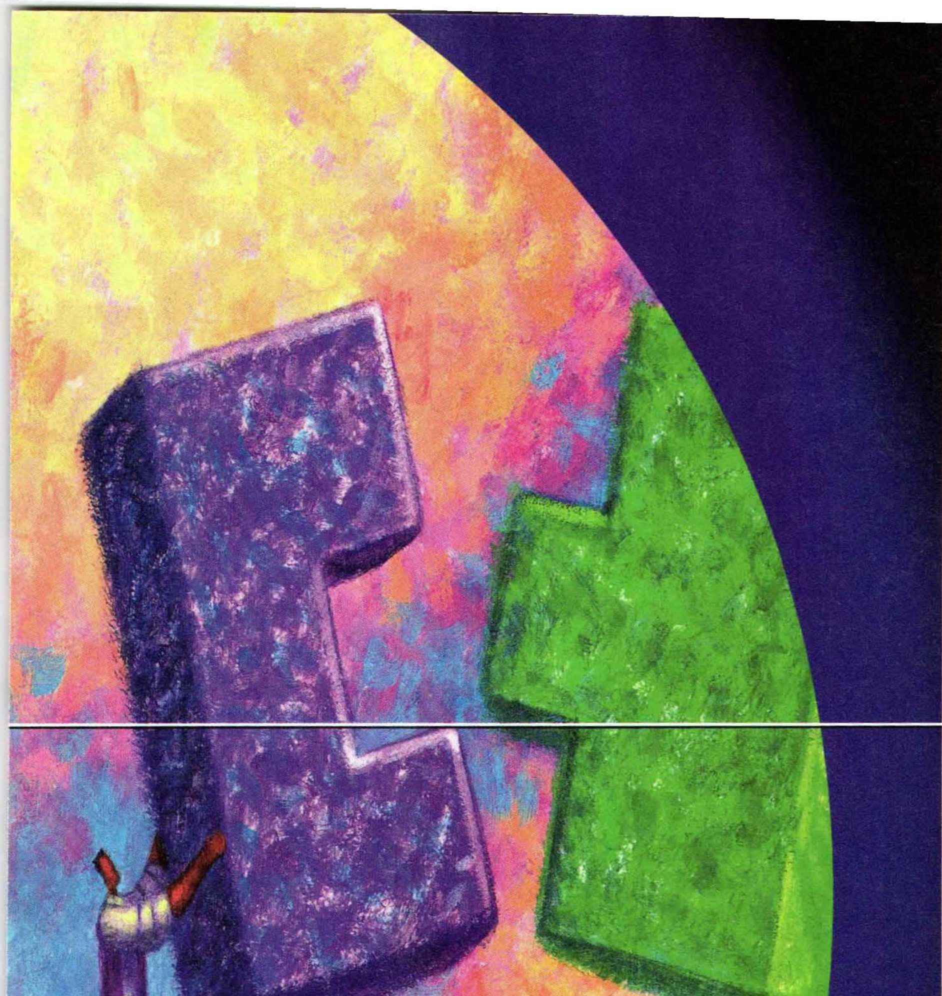

Wireless Show Promises Practical Design Solutions

Entering its 12th year, The Wireless Systems Design Conference & Expo continues to be a source of ideas and education for designers facing wireless design and measurement issues.

JACK BROWNE
Publisher/Editor

Wireless device, circuit, and system design engineers must face time-to-market pressure every day. Challenged also by difficult economic times, designers rely on the latest information on wireless design techniques, test equipment, and software tools in order to be effective and efficient. As it has for 11 years, the Wireless Systems Design Conference & Expo 2004 (formerly known as the Wireless Symposium & Exhibition) remains true to its charter of providing practicing engineers with practical design solutions for current problems, rather than theory and future-looking market studies. The event is scheduled for March 8-10, 2004 in the San Diego Convention Center (San Diego, CA). This second of three special reports previewing the 12th Annual Wireless Systems Design Conference & Expo will highlight some of the more than 60 technical sessions scheduled for San Diego.

As noted last month, Dr. Henry Samueli, co-founder, chairman, and chief technical officer (CTO) of Broadcom Corp. (Irvine, CA) will speak the first day of the conference on life in a wireless world and business trends related to the growth of wireless networks. The following day, a second keynote address, by Ron Reedy, founder, vice-president, and CTO of Peregrine Semiconductor Corp. (San Diego, CA), will illuminate some of the similarities between working for military customers compared to designing for wireless customers. Dr. Reedy, who founded Peregrine Semiconductor (www.peregrine-semi.com) in 1990, is no stranger to military markets, having been a branch head of the microelectronics division at Navy Research and Development (NRaD). Holding a BSEE from the US Naval Academy and MSEE from the US Naval Postgraduate School, he is now responsible for products and services delivered to commercial, military, and space customers. The company's unique Ultra-Thin Silicon-on-Sapphire (UTSi)® CMOS process offers the high levels of integration possible with CMOS and the good high-speed/high-frequency performance associated with GaAs semiconductors. (For more on the UTSi process, read the "Application Notes" section on p. 94.)

The San Diego Wireless Systems Design Conference & Expo 2004 features several presentations on a technology of great promise for high-data-rate wireless applications: ultrawideband (UWB) technology. For example, on Tuesday, March 9th, Jon Adams of Motorola SPS Wireless and Mobile Systems Group will discuss the fundamentals of UWB technology and how this short-pulsed, "carrierless" com-

munications technique can achieve data rates exceeding 100 Mb/s over short distances with low-voltage devices and mere milliamperes of current. That same day, Ian Oppermann, the Director of the Centre for Wireless Communications, will explore ad hoc multihop wireless networks based on UWB and other broadband technologies. As part of a technical track on broadband/wideband networks, his presentation will project future directions in wireless ad hoc multihop networks, focusing on technologies for low-power, short-range communications systems. His research combines three main fields: low-overhead Internet protocols (nanoIP), ad hoc networks operating with heterogeneous radio structures, and UWB technologies for positioning and communications functions. The presentation will include results from several European Union (EU) UWB tests and trials. Finally, Bohdan Stryzak of Photran Sciences offers a presentation comparing UWB and ultranarrowband (UNB) modulation formats, and how each modulation format provides advantages for different applications.

Traditional Technologies

For those seeking more traditional wireless technologies, the Wireless Systems Design Conference & Expo also features technical tracks on cellular and third-generation (3G) technologies, handset design strategies, low-power design, component design, and software development. In the Cellular/3G track, for example, Jim Person of the CDMA Development Group, highlights the growth of CDMA2000 technology and its use in wireless multimedia applications (currently more than 54 million CDMA2000 subscribers). His presentation will examine the migration to CDMA2000 from other standards and describe CDMA2000 deployment experiences by means of real-life examples.

Also in the Cellular/3G track, Michael Civiello of Zyray Wireless will explain how to make the transition to 3G technology through the design of low-cost, backward-compatible wideband CDMA

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(WCDMA) handsets. His approach involves the use of an add-on WCDMA baseband processor designed to interface with existing GSM/GPRS handset designs. In the same track, Grover Righter of Kabira Technologies outlines an approach for the deployment of real-time delivery of wireless content and messaging services (service on demand). He will explain how the newer software layers interact with a wireless network for messaging applications and how to link business-model requirements with design architectures in 2.5G and 3G networks. Also in the Cellular/3G track, Sanjeev Varma of Airvana will look beyond 3G and project the types of technologies that will be needed for next-generation wireless services.

The component design track will feature a diversity of topics ranging from memory and field-programmable gate arrays (FPGAs) to antennas and custom substrates. For example, Robert Markunas of Ziptronix will detail how to improve wireless system performance through the use of "engineered" substrates. These customized substrates can be used, in one instance, to minimize the temperature-related performance variations of surface-acoustic-wave (SAW) filters commonly used in wireless designs. In the same session, Jaekyun Moon of Bermai shows how antenna diversity can be used for improved OFDM wireless performance.

Power Management

In the power-management track, Peter Henry of National Semiconductor reports on advances in portable power. Pointing to the increased integration of wireless functions, such as positioning, messaging, video, and voice, within a single portable platform, Henry promises to review new technologies that are designed to increase battery life expectations. In the same session, Derek Koonce of Vishay Siliconix details the challenge of improving the efficiency of DC-to-DC conversion, hopefully through the use of novel P-channel MOSFET devices that can improve power-supply efficiency. Also in the power-management session, Jim Wight of IceFyre Semiconductor reveals a new power amplifier design that sets new standards of efficiency for 5-GHz orthogonal-frequency-division-multiplexing (OFDM) wireless-local-area-network (WLAN) systems. The unique switched-mode amplifier employs low-loss Chireix power combiners to achieve high levels of output power with low voltage and current.

Next month's report will continue the review of the more than 60 technical presentations scheduled for the Wireless Systems Design Conference & Expo, along with a summary of key exhibitors and their latest products. For more information on the 12th Annual Wireless Systems Design Conference & Exposition, visit the website at www.wsdexpo.com. MRF

Measurement Group Meets On Test/Modeling Issues

The most recent meeting of the ARMMS RF and Microwave Society featured technical presentations on optimum use of CAE simulation tools and precision test equipment.

JACK BROWNE
Publisher/Editor

Simulation and measurement results should agree closely under ideal conditions. Although such conditions are often not met, the most recent meeting of the ARMMS RF and Microwave Society Meeting focused on coordinating design efforts with computer-aided-engineering tools and accurate measurement techniques with microwave test equipment for the improvement of both commercial and military designs. Coordinated by Roger Hopper of Roke Manor Research Ltd. (Roke Manor, Romsey, Hamshire, England), the ARMMS RF and Microwave Society Meeting was held November 3-4, 2003 in the Hotel Elizabeth (Rockingham, Corby, Northamptonshire, England).

The two-day meeting offered modeling and measurement solutions for both commercial and military applications. The opening report, for example, by Malcolm Edwards of AWR Ltd. (Hitchin, Hertsfordshire, England), reviewed the differences among the major wireless-local-area-network (WLAN) standards, such as IEEE 802.11a/b/g, and some of the approaches used to design WLAN radios, including as integrated multichip modules and single-chip receivers. His presentation included a discussion on orthogonal frequency division multiplexing (OFDM), which is designed to reduce crosstalk between channels and minimize the effect of multipath distortion, and important parameters for testing WLANs, such as adjacent-channel-ratio (ACPR) and error-vector magnitude (EVM).

Edwards' presentation was followed by Kelvin Clarke of Ansoft UK who explained how many complex design problems could be broken down into constituent parts. This "divide and conquer" strategy can lead to very efficient and accurate solutions. The basic approach involves deciding how to best subdivide large problems and solve the constituent problems parametrically and then construct models from constituent parts. The availability of parametric three-dimensional (3D) models enables fast design and optimization of very large structures. Software based on finite-element (FE) analysis solves Maxwell's equations for the volume of arbitrary 3D structures. The FE method can also be applied to solve transfinite-element problems to determine the two-dimensional field solution at a given port of a "black box" design.

Chris Mann of Flann Microwave Ltd. (Bodmin, Cornwall, England) spoke on using the HFSS 3D field solver from Ansoft Corp. (Pittsburgh, PA), his report was inspired by the high cost of fabricating millimeter-wave waveguide components. HFSS provides an engineering environment in which engineers can experiment without cutting metal or operating expen-


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sive computer-controlled machinery. The report offered several design examples, including modification of a broadband dual polar horn antenna that operates from 2.5 to 18.0 GHz to cover a new frequency range of 18 to 40 GHz. Eventually, the software was used to optimize the design for a frequency range of 6 to 50 GHz. The software was also used to improve a 59-to-64-GHz omnidirectional antenna for WLAN use in the unlicensed 60-GHz band and to optimize an extremely compact Gaussian antenna.

Well-known design consultant, lecturer, and amplifier guru Steve Cripps offered a report on a high-efficiency Class F amplifier using a Chireix power combiner. Amidst lectures on advanced computer-aided-engineering (CAE) tools, Cripps explained how the amplifier, designed for a WLAN chip set from IceFyre (Kanata, Ontario, Canada), was created as a breadboard design built on hand-milled FR4 Test results revealed efficiency of better than 70 percent.

John Birkbeck of Roke Manor Research Ltd. covered issues associated with the design of RF and DC circuits for a custom heterojunction-bipolar-transistor (HBT) MMIC amplifier in parallel with packaging and thermal design. Design approaches must consider reducing the amplifier's sensitivity to process, temperature, and supply rail variations.

Nick Long of Great Circle Design (Somerset, England) explored a technique for detecting and measuring imperfections in I/Q signal processing systems. The technique can be used for fault analysis, design evaluation, and for alignment and calibration of systems.

Guy Purchon of Anritsu Ltd., European Measurement Division (Rutherford Close, Stevenage, Hertsfordshire, England) described the design and measurement principles for a new-high speed dual display channel power meter capable of detecting and displaying modulated signals. He explained how the tool could be used for commercial (WCDMA, WLAN) and military (radar) applications.

Kevin Lees and Stephen Protheroe of National Physics Laboratory (NPL, Teddington, England) discussed a traceable radiometer for calibration of noise sources below 10 MHz. They showed typical calibration results for a commercial noise source at 0.784 MHz and noted that their lowest traceable frequency was 100 kHz.

The next ARMMS meeting is scheduled March 29-30, 2004 at the Milton Hill House (Milton Hill, Steventon, Oxfordshire, England). For more information about ARMMS, visit the web site at www.armms.org, or contact JJ Heath-Caldwell, marketing coordinator, at 01962-761-565 (e-mail: jj@jjhc.co.uk) or Duncan McIntosh, secretary (e-mail: duncanmcintosh@ieee.org). MRF

Amplifiers Boost Signals To 31 GHz

THREE MEDIUM-POWER GaAs PHEMT amplifiers have been added to a line already spanning 12 to 32 GHz. Model HMC490 is a low-noise amplifier (LNA) that operates from 12 to 17 GHz with 26 dB gain and 2.2 dB noise figure. It achieves +26 dBm output power at 1-dB compression with a third-order intercept point of +35 dBm. Model HMC498 provides 24 dB gain and +25 dBm output power at 1-dB compression from 17 to 24 GHz. It offers +27 dBm saturated output power with a third-order intercept point of +34 dBm. Model HMC499 provides 16 dB gain and +25 dBm saturated output power from 21 to 32 GHz with 25-percent power-added efficiency. It offers 45 dB reverse isolation with third-order intercept point of +33 dBm.

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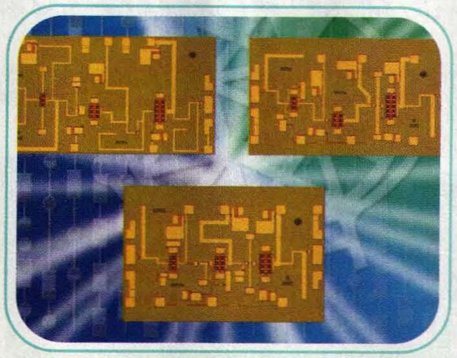
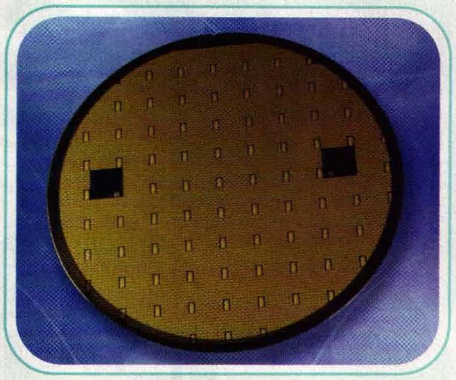


HITTITE'S AMPLIFIERS

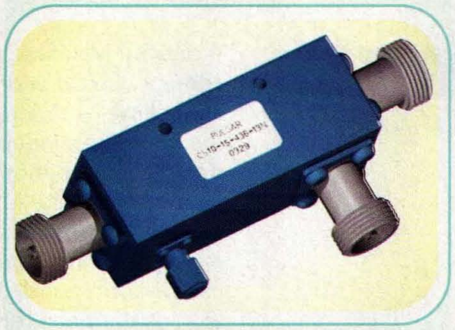


TRIQUINT'S TQPED E-D PHEMT PROCESS

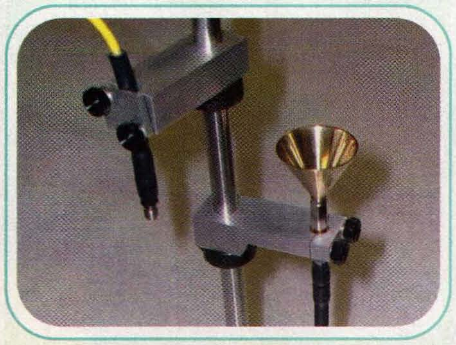
GaAs Process Offers High-Level Integration

INCLUDING BOTH enhancement-mode and depletion-mode PHEMTs, the TQPED E-D PHEMT process combines high-frequency active devices with on-chip passive elements, three layers of interconnections, and substrate viaholes for higher levels of integration on MMICs than previously possible with single-mode transistor processes. The new TQPED process offers enhancement-mode transistors with cutoff frequency of 25 GHz and maximum frequency of oscillation of 100 GHz. It provides depletion-mode transistors with cutoff frequency of 25 GHz and maximum frequency of oscillation of 90 GHz. The process also features MIM capacitors and NiCr and bulk resistors with 50 and 285 ohms/square, respectively. The high-volume process is ideal for integrating analog and control functions for a wide range of wireless applications.

TriQuint Semiconductor, 2300 NE Brookwood Pkwy., Hillsboro, OR 97124; (503) 615-9000, FAX: (503) 615-8900, Internet: www.triquint.com.



PULSAR'S COMPACT DIRECTIONAL COUPLERS



DAMASKOS' MODEL M035T COAXIAL CELL SETUP

Directional Couplers Span 2 To 18 GHz

A LINE OF compact directional couplers with Type N coaxial connectors has been developed for applications from 2 to 18 GHz, including in military and commercial communications systems and for general-purpose measurement setups. Models CS8-15-436, CS10-15-436, and CS20-15-436 are available with coupling values of 8, 10, and 20 dB, respectively. The coupling flatness for each unit is ±1 dB, while the insertion loss is 0.8 dB or less. These broadband coaxial couplers feature frequency response that is within ±0.5 dB with maximum VSWR of 1.50:1. The directivity is 15 dB from 2 to 12 GHz and 12 dB from 12 to 18 GHz.

Pulsar Microwave Corp., 48 Industrial Way, Clifton, NJ 07012; (800) 752-2790, (973) 779-6262, FAX: (973) 779-2727, e-mail: sales@pulsarmicrowave.com, Internet: www.pulsarmicrowave.com.

Set Tests Liquid/Powder Dielectric Constants

THE MODEL M035T 3.5-mm Coaxial Cell Setup measures the complex dielectric constant and permeability of liquids and powdered materials at frequencies to 26 GHz. The system is supplied with calibration standards, loading aids, auxiliary equipment, and the company's MU-EPSLN™ software. In addition to controlling the analysis equipment, such as a microwave vector network analyzer (support is provided for instruments from Anritsu Company and Agilent Technologies), the software coordinates the measurement of S-parameters, calibrates coaxial and waveguide lines, processes data, and calculates the values of dielectric constant and permeability for liquids and powders under test. The compact measurement cell avoids the problems associated with contact-probe methods and provides the full permeability spectrum of the material under test.

Damaskos, Inc., P.O. Box 469, Concordville, PA 19331; (610) 358-0200, FAX: (610) 558-1019, Internet: www.damaskosinc.com.

GPS Market Revenue Is Growing

AS CONSUMER AWARENESS of Global Positioning Systems (GPS) increases, so has product innovation and total market revenue. While roughly half of the market today consists of sales of automotive and asset-tracking equipment, these segments will still continue to grow at rates faster than that of the broader market for GPS equipment. Despite the strength of these markets, new segments are constantly emerging for GPS applications, driving demand for gear as diverse as people-tracking devices and GPS golf systems. The net result will be a market worth over $22 billion by 2008, according to Oyster Bay, NY-based technology market-research firm ABI.

Companies like Garmin, Wherify Wireless, and Navman are synonymous with integrating GPS receivers into innovative form factors. Advances in GPS integrated circuits (ICs) will fuel this trend across the entire industry. Sony's recent announcement unveiling a miniature, single-chip IC provides further evidence that more of these novel applications are likely in an ever-increasing range of devices.

According to a recent study from ABI, unit growth in ICs, the brains behind the devices' positioning capabilities, will likely be at about 35-percent compounded annually over the next five years. Revenue growth will be strong, but not as spectacular as unit growth, due mainly to pricing pressure.

The study, "GPS World Markets: Opportunities for Equipment and IC Suppliers," examines the current status and trends of the GPS industry. Covered areas include wireless and in-vehicle navigation, as well as growing segments such as recreation, communication, people tracking, marine and surveying, among others. For each segment, total market value is forecasted to 2008 in addition to regional totals. An analysis of key market drivers and barriers for each segment is presented. The report also quantifies the market for GPS IC shipments, ASP, and revenue to 2008.

Further information on this subject can be found on ABI's website at www.abiresearch.com. MRF

CONTRACTS

Raytheon Co.—Has been awarded an additional $59 million to its Global Broadcast Service (GBS) contract by the US Air Force's Electronic Systems Center at Hanscom AFB in Massachusetts, extending services through December 2005. Raytheon will perform the work at various customer sites with management and administrative support out of its Reston, VA facility.

Raytheon's work on the contract will focus on engineering, production, logistics, and operational support to upgrade the existing Air Force GBS satellite communications, deployed worldwide, to all of the US military services. The system, operational after the events of September 11, 2001, provides high-capacity broadcast products, including audio, video, files, and Web services.

Reynolds Industries—Announced that Rockwell Collins of San Jose, CA has awarded Reynolds a contract for helmet-vehicle-interface (HVI) cable assemblies for the Joint Helmet Mounted Cueing System (JHMCS) Program. This award is for the LRIP-4 phase.

FRESH STARTS

Hittite Microwave Corp.—Announced the appointment of a new sales representative firm to serve customers in Northern Germany. MRC Components OHG, headquartered in Freising, Germany, was established in 1997 to act as a specialist European representative for electronic component manufacturers from the USA. MRC Components' areas of expertise include: RF and microwave components; fiber-optic components; power supplies; and magnetic core materials and magnetic components.

Tessera Technologies, Inc.—Announced that it has priced its initial public offering of 7.5 million shares of its common stock at $13.00 per share. Of these 7.5 million shares, the company will issue 3.0 million shares and existing stockholders will sell an additional 4.5 million shares in the offering. In addition, the company and the selling stockholders have granted the underwriters an option to purchase up to an additional 1,125,000 shares to cover over-allotments, if any.

LPKF Laser & Electronics AG—Has announced that they have signed a corporation agreement with Harting Technology Group (Espelkamp, Germany). This partnership allows Harting to produce Three Dimensional Molded Interconnect Devices (3D-MID's) for the micro packaging field, using LPKF's proprietary Laser Direct Structuring (LDS) process.

With this agreement, Harting, a supplier of electrical, electronic, and optical connectors and components, will become a full-featured supplier of LDS made 3D-MID solutions.

Elcoteq Network Corp.—Has moved its Americas headquarters to a new facility in Irving, TX. Growth in sales and services and an increase in personnel necessitated more space. The new offices are 60-percent larger than the previous ones and contain the executive staff, administration, IT, marketing, sales, human resources, and supply-chain management. Elcoteq Americas includes manufacturing and NPI facilities in Monterey, Mexico and Dallas, TX, and employs 1050 people.

The new contact information is: 909 Lake Carolyn Pkwy., Suite 500, Irving, TX 75039; (972) 401-9995, FAX: (972) 401-9606.

CableLabs—Has moved to a new location. The new contact information is: Cable Television Laboratories, Inc., 858 Coal Creek Circle, Louisville, CO 80027-9750; (303) 661-9100, FAX: (303) 661-9199.

The Summation Research Family of Companies—Announced the formation of SRI Hermetics, Inc. This undertaking will enable the family of companies to expand the hermetic connector business of SRI Connector Gage Co. as well as offer new product capabilities to Summation Research, Inc. and SRI PMD, Inc.

SRI Hermetics will be headquartered with the other Summation Research companies in Melbourne, FL.

TriQuint Semiconductor, Inc.—Received its first purchase orders from a major Korean handset manufacturer for its Model 890035 triplexer—an integrated, front-end solution for CDMA wireless phones. A triplexer is a passive filter module that allows the simultaneous use of a single antenna for both CDMA and GPS bands, reducing component count and design complexity while freeing handset board space.

Semflex, Inc.—Announced the appointment of a new sales rep organization, Tech Marketing Associates, to serve the Southern California region.

Headquartered in Costa Mesa, CA and managed by Frank Jones, the company can be reached at (714) 979-0412 or by e-mail at jonesfc@aol.com. For further information about Semflex products, call Semflex Customer Service at (800) 778-4401 or visit the website at www.semflex.com.

CBC/Radio-Canada and SIRIUS—Will form a joint venture to bring satellite radio to Canada. CBC/Radio-Canada and SIRIUS also announced that the venture will soon file an application with the Canadian Radio-television and Telecommunications Commission (CRTC) for a license to provide satellite radio in Canada.

Silicon Laboratories, Inc.—Has completed the acquisition of privately held Cygnal Integrated Products, an Austin, TX firm involved in the microcontrollers business.

In connection with the acquisition, Silicon Laboratories issued approximately 1.2 million shares of common stock in exchange for all outstanding shares of Cygnal capital stock. Up to approximately 1.3 million additional shares of Silicon Laboratories' common stock will be reserved for future issuance to the shareholders of Cygnal based on the achievement of certain revenue milestones following the closing of the transaction. MRF

THE GLOBAL SOLUTION... AND BEYOND!

10MHz to 7GHz AMPLIFIERS from $99.95 (1-9 qty.)

From amateur radio to cellular to satellite applications, with medium output power up to 17dBm, Mini-Circuits versatile ZJL and ZKL connectorized amplifiers offer the broad range of choices designers demand for achieving high system performance goals. Ultra-wideband models deliver **gain ranging from 9 to 40dB** and IP3 up to +32dBm. But beyond the performance and reliability built into these miniature 12V amplifiers lies another important feature, the low price...from only $99.95! Call now for fast delivery.

Mini-Circuits...we're redefining what VALUE is all about!

SPECIFICATIONS

Model	Freq (MHz)	Gain (typ) Midband (dB)	Gain (typ) Flat (±dB)	Max. P_{out}[1] (dBm)	Dynamic Range (Typ @2GHz[2]) NF(dB)	Dynamic Range (Typ @2GHz[2]) IP3(dBm)	I(mA)[3]	Price $ea. (1-9)
ZJL-5G	20-5000	9.0	±0.55	15.0	8.5	32.0	80	129.95
ZJL-7G	20-7000	10.0	±1.0	8.0	5.0	24.0	50	99.95
ZJL-4G	20-4000	12.4	±0.25	13.5	5.5	30.5	75	129.95
ZJL-6G	20-6000	13.0	±1.6	9.0	4.5	24.0	50	114.95
ZJL-4HG	20-4000	17.0	±1.5	15.0	4.5	30.5	75	129.95
ZJL-3G	20-3000	19.0	±2.2	8.0	3.8	22.0	45	114.95
ZKL-2R7	10-2700	24.0	±0.7	13.0	5.0	30.0	120	149.95
ZKL-2R5	10-2500	30.0	±1.5	15.0	5.0	31.0	120	149.95
ZKL-2	10-2000	33.5	±1.0	15.0	4.0	31.0	120	149.95
ZKL-1R5	10-1500	40.0	±1.2	15.0	3.0	31.0	115	149.95

NOTES:
1. Typical at 1dB compression.
2. ZKL dynamic range specified at 1GHz.
3. All units at 12V DC.

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For quick access to product information see MINI-CIRCUITS CATALOG & WEB SITE • EEM • MICROWAVE PRODUCT DATA DIRECTORY • WWW.RFGLOBALNET.COM

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BYRNES

Andrew Appoints Byrnes As VP/GM For SatCom

Andrew Corp. has named JOAN M. BYRNES to fill the newly created role of vice president and general manager for satellite communications. Byrnes was previously the COO at Loral Skynet, a division of Loral Space & Communications.

SiGe Semiconductor—BOB LITZLBECK to director of sales for North America; formerly director of wireless sales at Infineon Technologies.

TT electronics—ALLEN HILTON to vice president and general manager of the IRC Wirewound and Film Technologies Division; formerly director of operations with IRC's sister company, BI Technologies, in Mexicali, Mexico, Kuala Lampur, Malaysia, and Fullerton, CA. Also, ALLAN COLE to director of sales and marketing; formerly CEO at Yaego.

Enthone, Inc.—RAYMOND FONG to vice president and managing director for the Asia-Pacific region; formerly served as the Asia-Pacific general manager for Avery Dennison Ltd.

Rosenberger LEONI Site Solutions—BILL CORONDAN to the North American sales initiative; formerly sales manager with the Broadband Sales Group at Andrew. Also, RICK HARRIS to the North American sales initiative; formerly vice president of sales for Sky Cross.

Leitch Technology Corp.—TIMOTHY E. "TIM" THORSTEINSON to president and CEO; previously headed the Broadcast product line at Thomson Broadcast and Media Solutions.

Southwest Research Institute (SwRI)—JAMES ROBERT (BOB) KEYS to director of the newly formed Applied Power Department of SwRI's Applied Physics Division; formerly assistant director of SwRI's Electronics and Power Systems Department.

Texas Instruments—CHRISTINE TODD WHITMAN to the board of directors; formerly administrator of the Environmental Protection Agency (EPA) and Governor of the state of New Jersey.

California Micro Devices—DR. DAVID W. SEAR to the board of directors; formerly president and CEO of Vaishali Semiconductor and Integrated Circuit Systems.

Tyco International Ltd.—DAVID POLK to vice president of media relations; formerly vice president of communications for Raytheon's Integrated Defense Systems business.

Park Electrochemical Corp.—KONSTANTINE N. ("GUS") KARAVAKIS to director of research and development; formerly director of technology for Nexcleon, Inc.

Speedline Technologies, Inc.—DENNIS O'NEAL to director of materials deposition products; formerly product manager for semi-automatic printers. Also, MARC APELL to director of thermal and cleaning products; formerly reflow product manager.

Janos Technology, Inc.—FREDERICK MULLAVEY to director of quality; formerly Six Sigma Black Belt Consultant for Quest Telecommunications in Phoenix, AZ, through A&A Services (Stone Mountain, GA).


MULLAVEY


GANDENBERGER

Micrel Inc.—JAMES GUY GANDENBERGER to vice president of Wafer Fab Operations; formerly managing director of Santa Clara Wafer Fabs at National Semiconductor. MRF

▸SHORT COURSES

Optical Fibre Measurements
February 2-4 (Teddington, Middlesex, UK)
National Physical Laboratory
For further information, contact:
Martin Wicks
020 8943 6787, FAX: 020 8943 6098
e-mail: martin.wicks@npl.co.uk
Internet: www.npl.co.uk/photonics/training/ofmc2004.html

UCLA Extension Short Courses
All courses at UCLA Extension Building, Los Angeles, CA
February 9-13 *Satellite Communications Networks and Applications: Selecting the Right Technology and Methodology*
February 23-26 *Space-Based Radar: Techniques and Design*
February 25-27 *Satellite Thermal Control*
For further information, contact:
Dr. William R. Goodin
UCLA Extension, Short Course Program
10995 Le Conte Ave., Suite 542
Los Angeles, CA 90024-2883
(310) 825-5010, FAX: (310) 206-2815
e-mail: bgoodin@unex.ucla.edu
Internet: www.uclaextension.edu/shortcourses

Technology Strategy for R&D and Product Development
March 25-26 (Pasadena, CA)
Course Location: California Institute of Technology Industrial Relations Center
For further information, contact:
California Institute of Technology
Industrial Relations Center, I-90
Pasadena, CA 91125-9000
(626) 395-4043, FAX: (626) 795-7174
e-mail: execedu@caltech.edu
Internet: www.irc.caltech.edu

Antenna Parameter Measurements By Near-Field Techniques
April 20-22 (Boulder, CO)
National Institute of Standards and Technology (NIST) Boulder Laboratories
To register, contact:
Wendy McBride, Conference Manager
NIST, MS 346
325 Broadway
Boulder, CO 80305
(303) 497-4500, FAX: (303) 497-5208
e-mail: Wmcbride@boulder.nist.gov
Internet: www.nist.gov/public_affairs/confpage/blconf.htm

▸MEETINGS

CONMET Conference & Expo
January 26-29 (Washington, DC)
Washington Convention Center
Conference is January 26-29
Expo is January 27-29
(866) COMNETX
Internet: www.comnetexpo.com

ASQ's Six Sigma Conference 2004
February 2-3 (Phoenix, AZ)
Pointe Hilton Squaw Peak Resort
Internet: http://sixsigma.asq.org

Wireless Systems Design Conference and Expo 2004 co-located with the Military Electronics Show West 2004
March 8-10 (San Diego, CA)
San Diego Convention Center
To exhibit, contact Sharon Pierce at (203) 559-2968 or spierce@penton.com
For marketing opportunities, contact Amy Orsini at (203) 559-2966 or aorsini@penton.com
FAX: (203) 559-2840
Internet: www.wsdexpo.com

CeBit 2004
March 18-24 (Hannover, Germany)
(609) 987-1202 ext. 211
e-mail: cebit@hfusa.com
Internet: www.hfusa.com/cebit

electronicaUSA with the Embedded Systems Conference
March 29-April 1 (San Francisco, CA)
Moscone Convention Center
For further information, contact:
Linsy Reese, CMP Media LLC
(415) 947-6645, FAX: (415) 947-6009
e-mail: lreese@cmp.com
Internet: www.electronicaUSA.com

RF & Hyper Europe 2004
March 30-31, April 1 (Paris, France)
Paris Expo—Porte De Versailles—Hall 2.2
For further information, contact:
BIRP
11, rue du Perche
75003 Paris, France
+33 1 44 78 99 30, FAX: +33 1 44 78 99 49
e-mail: hyper@birp.fr
Internet: www.birp.com/hyper

2004 International Conference on Compound Semiconductor Manufacturing Technology
May 3-6 (Miami Beach, FL)
Sheraton Bal Harbour Beach Resort
e-mail: info@gaasmantech.org
Internet: www.gaasmantech.org or www.csmantech.org

2004 Symposium on VLSI Technology
June 15-17 (Honolulu, HI)
Hilton Hawaiian Village
e-mail: vlsi@vlsisymposium.org
Internet: www.vlsisymposium.org

▸CALL FOR PAPERS

The 5th Topical Meeting on Silicon Monolithic Integrated Circuits in RF Systems
September 8-10 (Atlanta, GA)
Georgia Tech
Submission deadline: March 1
Status notification: April 1
Final manuscripts due: June 1
For conference details and paper submission information, contact Chris Evans, Conference Coordinator, at sirf04@ece.gatech.edu
Internet: www.ece.gatech.edu/conferences/sirf04

Analyzer Effects Of DSSS On Analog FM Broadcasts

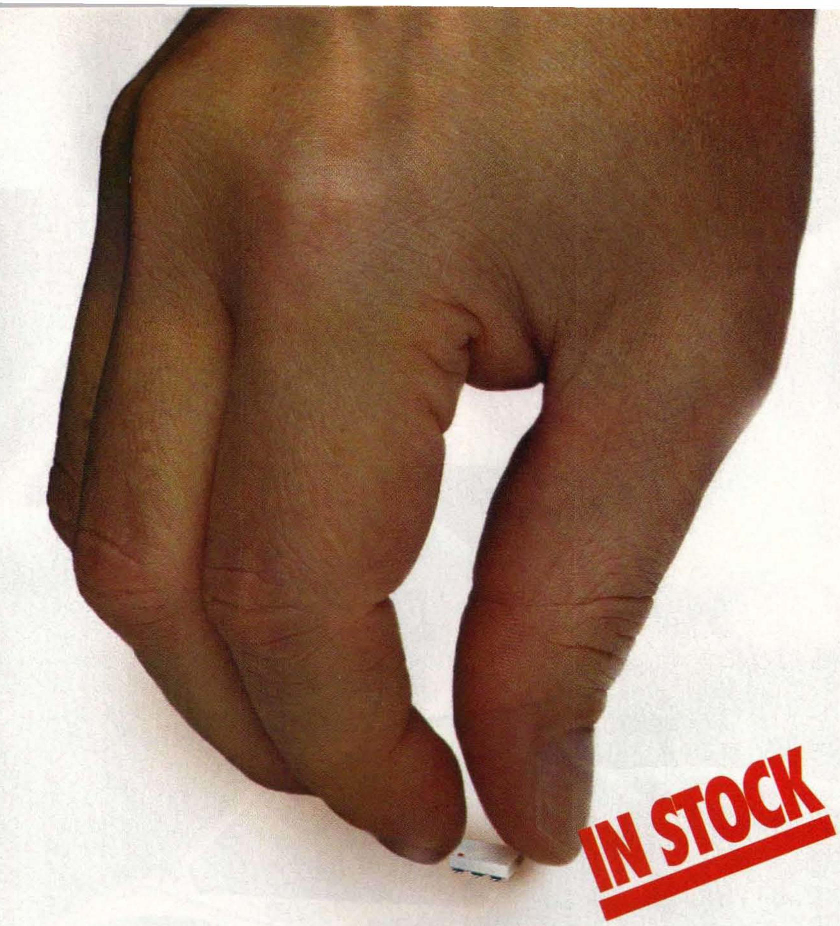
SPREAD-SPECTRUM COMMUNICATIONS systems are noteworthy for their capability of coexisting with other communications formats, including frequency-modulation (FM) broadcast systems. But does the presence of a spread-spectrum system interfere in any way with an existing analog or digital communications systems? Demosthenes Vouyioukas and Philip Constantinou of the Department of Electrical and Computer Engineering of the National Technical University of Athens (Athens, Greece) sought the answer to this question through their research on a direct-sequence-spread-spectrum (DSSS) system overlaid on an existing analog FM broadcast system, with both systems occupying the same frequency band. The basic goal of the research was to find the amount of spread-spectrum interference that could be tolerated by an analog FM broadcast system without excessive degradation in FM broadcast audio quality. The standard FM system operates from 87.5 to 108.0 MHz with 200-kHz channel spacing and peak frequency deviation of 75 kHz for the channels. The DSSS system is characterized by a power spectral density that closely resembles additive white Gaussian noise (AWGN) within the same bandwidth. Laboratory tests were performed to determine the performance of the analog FM signal in the presence of co-channel and adjacent-channel interference from the DSSS signals. In general, the results show that the spread-spectrum interference causes degradation of the analog FM baseband signals by increasing the noise and therefore decreasing the signal-to-noise ratio (SNR) of the FM receiver used in the tests. As the spreading of the DSSS signals increased, the effects of the SNR degradation decreased. See "Performance Degradation of Analog FM System Due to Spread Spectrum Overlay," *IEEE Transactions on Broadcasting*, June 2003, Vol. 49, No. 2, p. 113.

Branch-Line Coupler Aids Design Of CPW Balanced Amp

WIRELESS LOCAL-AREA NETWORKS (WLANs) continue to spread throughout businesses and small offices, creating a need for high-performance amplifiers for improved WLAN coverage. To fill this need, Bub-Sang Yun and associates from the RFIC Research and Education Center & Mission Technology Research Center at Kwangwoon University (Seoul, Korea) have developed an efficient coplanar-waveguide (CPW) balanced amplifier with a branch-line coupler to combine the power stages. Using advanced computer-aided-design (CAD) tools from Agilent Technologies (Santa Rosa, CA) and Zeland Software (Fremont, CA), the researchers chose the CPW structure due to its relative insensitivity to viahole/substrate thickness compared to traditional microstrip approaches. The CPW technology offers simplicity of fabrication and eliminates the need for backside processing (and production cost). The researchers directly printed the amplifier circuitry on Duroid 6010 substrate material from Rogers Corp. (Rogers, CT). The single-stage amplifier, which is designed for use as both a WLAN power amplifier or low-noise amplifier (LNA), was tested with an 8510C vector network analyzer from Agilent and found to have about 10-dB gain from 5.5 to 6.0 GHz with input return loss of about 16 dB and output return loss of 17.5 dB. The amplifier, which is based on a model ATF-13336 GaAs FET device from Agilent Technologies, draws about 25 mA current from a +2.5-VDC supply. See "CPW Balanced Amplifier With A New Branch-Line Coupler," *Microwave and Optical Technology Letters*, December 5, 2003, Vol. 39, No. 5, p. 349

Antennas Support Satcom With Moving Vehicles

SATELLITE COMMUNICATIONS has always faced the challenge of achieving effective links between geostationary satellites and moving vehicles on earth. The usual solution requires that a vehicle-borne antenna be circularly polarized and maintain a gain of at least 3 dBi in the direction of the satellite regardless of the orientation of the vehicle or its location within a latitude belt of significant breadth. A key to implementing the solution, of course, is that it be fairly simple and low cost to manufacture. F. Ares and associates from the University Santiago de Compostela (Coruna, Spain), the University of Naples (Naples, Italy), and the LEMA Swiss Federal Institute of Technology (Lausanne, Switzerland) developed two practical approaches including an array of circular patches and an array with square antenna patches, with two feeds per patch for circular polarization. See "Satellite Communication With Moving Vehicles On Earth: Two Prototype Circular Array Antennas," *Microwave and Optical Technology Letters*, October 5, 2003, Vol. 39, No. 1, p. 14.



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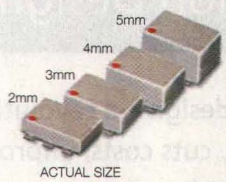
Searching high and low for a better frequency mixer? Then take a closer look at the Innovative Technology built into Mini-Circuits ADE mixers. ***Smaller size*** is achieved using an ultra-slim, patented package with a profile as low as 0.082 inches (2mm) in height. Electrically, ADE mixers deliver ***better performance*** than previous generation mixers through all welded connections with unique assembly construction which reduces parasitic inductance. The result is dramatically improved high frequency and IP2-IP3 performance. Plus, ADE's innovative package design allows water wash to drain and eliminates the possibility of residue entrapment. Another ADE high point is the ***lower cost***...priced from only $1.99 each. So, if you've been searching high and low for a mixer to exceed expectations...ADE is *it*™



ADE Mixers...Innovations Without Traditional Limitations!

ADE* TYPICAL SPECIFICATIONS:

MODEL	LO Power (dBm)	Freq. (MHz)	Conv. Loss Midband (dB)	L-R Isol. Midband (dB)	IP3 @Midband (dBm)	Height (mm)	Price ($ea.) Qty. 10-49
ADE-1L	+3	2-500	5.2	55	16	3	3.95
ADE-3L	+3	0.2-400	5.3	47	10	4	4.25
ADEX-10L	+4	10-1000	7.2	60	16	3	2.95
ADE-1	+7	0.5-500	5.0	55	15	4	1.99▲
ADE-1ASK	+7	2-600	5.3	50	16	3	3.95
ADE-2	+7	5-1000	6.67	47	20	3	1.99▲
ADE-2ASK	+7	1-1000	5.4	45	12	3	4.25
ADE-6	+7	0.05-250	4.6	40	10	5	4.95
ADEX-10	+7	10-1000	6.8	60	16	3	2.95
ADE-12	+7	50-1000	7.0	35	17	2	2.95
ADE-4	+7	200-1000	6.8	53	15	3	4.25
ADE-14	+7	800-1000	7.4	32	17	2	3.25
ADE-901	+7	800-1000	5.9	32	13	3	2.95
ADE-5	+7	5-1500	6.6	40	15	3	3.45
ADE-5X	+7	5-1500	6.2	33	8	3	2.95
ADE-13	+7	50-1600	8.1	40	11	2	3.10
ADE-11X	+7	10-2000	7.1	36	9	3	1.99▲
ADE-20	+7	1500-2000	5.4	31	14	3	4.95
ADE-18	+7	1700-2500	4.9	27	10	3	3.45
ADE-3GL	+7	2100-2600	6.0	34	17	2	4.95
ADE-3G	+7	2300-2700	5.6	36	13	3	3.45
ADE-28	+7	1500-2800	5.1	30	8	3	5.95
ADE-30	+7	200-3000	4.5	35	14	3	6.95
ADE-32	+7	2500-3200	5.4	29	15	3	6.95
ADE-35	+7	1600-3500	6.3	25	11	3	4.95
ADE-18W	+7	1750-3500	5.4	33	11	3	3.95
ADE-30W	+7	300-4000	6.8	35	12	3	8.95
ADE-1LH	+10	0.5-500	5.0	55	15	4	2.99
ADE-1LHW	+10	2-750	5.3	52	15	3	4.95
ADE-1MH	+13	2-500	5.2	50	17	3	5.95
ADE-1MHW	+13	0.5-600	5.2	53	17	4	6.45
ADE-10MH	+13	800-1000	7.0	34	26	4	6.95
ADE-12MH	+13	10-1200	6.3	45	22	3	6.45
ADE-25MH	+13	5-2500	6.9	34	18	3	6.95
ADE-35MH	+13	5-3500	6.9	33	18	3	9.95
ADE-42MH	+13	5-4200	7.5	29	17	3	14.95
ADE-1H	+17	0.5-500	5.3	52	23	4	4.95
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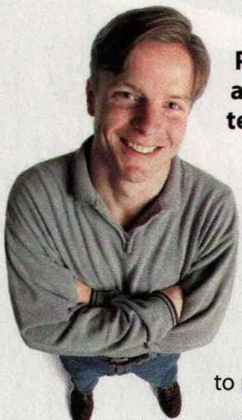

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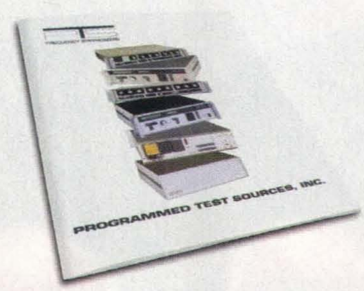

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Check Nonlinear Distortion With A Vector Signal Analyzer

Versatile vector signal analyzers can be used to study distortion on burst signals from base stations even while the mobile radio network is still operational.

Nonlinear distortion is a critical performance parameter for power amplifiers in mobile radio networks. Excessive distortion can degrade bit-error-rate (BER) performance, resulting in poor network voice and data transmission. Fortunately, the vector signal analyzer is an instrument designed not only for precise detection of vector and scalar modulation errors, such as error-vector-magnitude (EVM) characteristics, but also for the evaluation of amplifier and system-distortion characteristics. Since the analyzer does not require any special conditions or test signals for effective measurements, the instrument can be used to analyze burst signals from base stations while the mobile radio network is operational.

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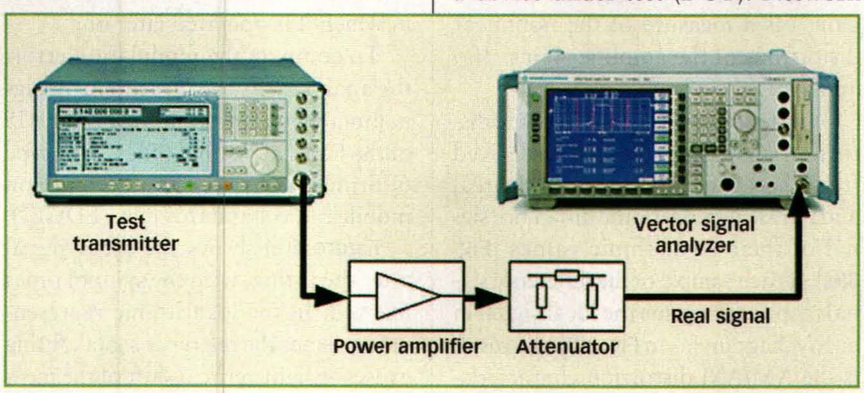
Traditionally, two-tone or multitone methods[1] were used with a selective voltmeter or a spectrum analyzer to determine the compression point of a device under test (DUT). Network analyzers allow similar approaches using power sweeps.[2] These methods employ test signals with little relation to "real-world" signals, or signals that are only optimized to the spectral bandwidth or the statistical signal distribution.

Vector signal analyzers are used to measure the scalar and vector modulation parameters and modulation errors of digitally modulated mobile radio signals. Modern concepts enable these instruments to be used for measuring and evaluating linearity errors as well, since all necessary data are collected during the course of normal measurements.[3] In fact, a standard test setup can be used without additional measuring instruments or special test signals.

Figure 1 shows a typical test setup for measurements using a vector signal analyzer. A signal generator with in-phase/quadrature (I/Q) modulation capabilities generates an RF mobile radio signal and applies it to the input of the DUT, for example a mobile radio output amplifier. The output of the amplifier is connected to the input of a vector signal analyzer (e.g., a model FSQ-K70 from Rohde & Schwarz) via



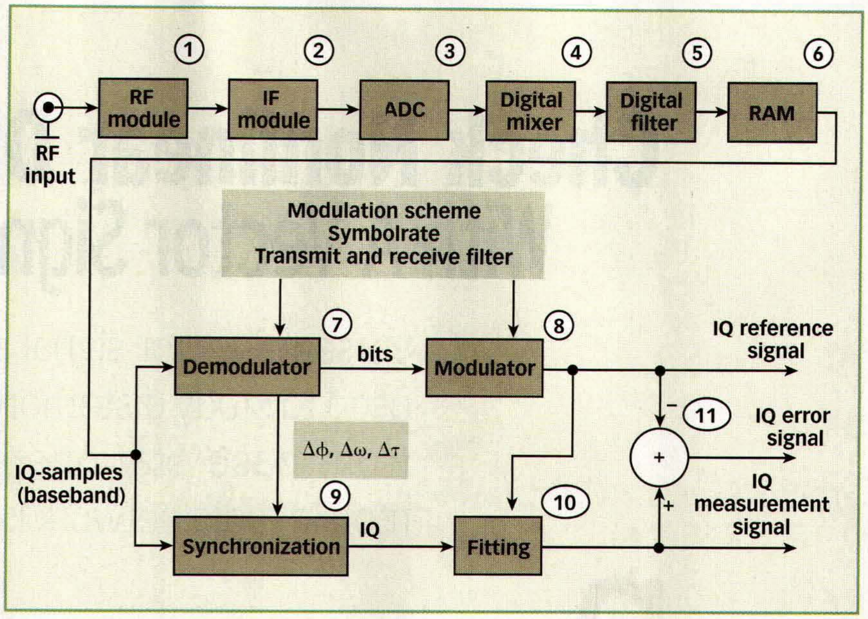
1. This basic test setup can be used for the measurement of nonlinear distortion in power amplifiers designed for mobile communications systems.

an attenuator (to avoid unacceptably high levels).This setup allows measurements even directly at the RF output of a base station.

Figure 2 shows a block diagram of a vector signal analyzer. The digitally modulated RF input signal passes through RF and intermediate-frequency (IF) stages (blocks 1 and 2 in Fig. 2) on its way to the input of the analog-to-digital converter [ADC](block 3 in Fig. 2). The IF signal is sampled, digitally mixed into a complex baseband signal (block 4 in Fig. 2), digitally filtered (block 5), and stored in random-access memory [RAM] (block 6).

A digital-signal processor (DSP) demodulates the baseband signals down to the bit level (block 7 in Fig. 2) and generates a "reference signal" corresponding to the undistorted transmit signal. The signal analyzer only needs to know the modulation scheme and appropriate filtering (block 8). After correcting for center-frequency offset, phase, and symbol timing (the synchronization block 9 in Fig. 2), the measured signal is fitted to the reference signal in magnitude and phase (block 10) in order to achieve a root-mean-square (RMS) value of the EVM. In the final stage, the measured signal and the reference signal are compared (block 11 of Fig. 2). Typical modulation errors, such as magnitude error versus time or phase error versus time are then calculated. These signals are used, for example, to display the vector and constellation diagrams or to subsequently calculate the distortion characteristics.

Figure 3(a) shows the ideal constellation diagram of an undistorted raised-cosine-filtered 16-state quadrature-amplitude-modulation (16QAM) signal. **Figure 3(b)** shows the output signal of an amplifier with pure amplitude distortion. Both figures display vector diagrams for complex baseband signals. The actual constellation points [Fig. 3(b)] are adjacent to their ideal positions. The curvature of the grid lines is a definite indication of nonlinear, modulation-dependent amplitude distortion. A section of the amplitude time characteristic can be seen in **Fig. 3(c).** The ideal signal appears as a blue curve while the real signal is shown as a red curve. Symbol times are marked by squares or circles for ease of identification. The three amplitude stages of the ideal signal are indicated by the horizontal lines R_1 to R_3, and those for the measurement signal are shown by the lines D_1 to D_3.



2. The various function blocks of a modern vector signal analyzer are shown in this high-level block diagram.

While the ideal and real signals still coincide at the lowest level stage, the deviations become larger with increasing level. If each level sample of the distorted signal and the corresponding sample of the ideal signal are plotted in an x/y diagram, the result is a modulation-dependent amplitude characteristic [**Fig. 3(d)**]. For better orientation, the level stages also appear as lines. The deviation of the characteristic from the diagonal ("linear gain") is a measure of the nonlinear distortions of the amplifier [Figs. 3(a) and 3(b)].

A practical representation of the distortion characteristic can be derived from the signal ratio between the real and the ideal signal or the difference signal of their logarithmic values [**Fig. 3(e)**]. If each sample of the difference signals is plotted versus the ideal signal in an x/y diagram [as in **Fig. 3(f)**], the result is the AM/AM distortion characteristic (amplitude-dependent amplitude distortion). All measurement points are used for interpolating the characteristic curve. In this representation, the deviation of the characteristic from the horizontal 0-dB line is a measure of the nonlinear distortion [Figs. 3(e) and 3(f)]. Likewise, the phase errors can be derived as a function of the ideal amplitude in an AM/PM characteristic (amplitude-dependent phase distortion).

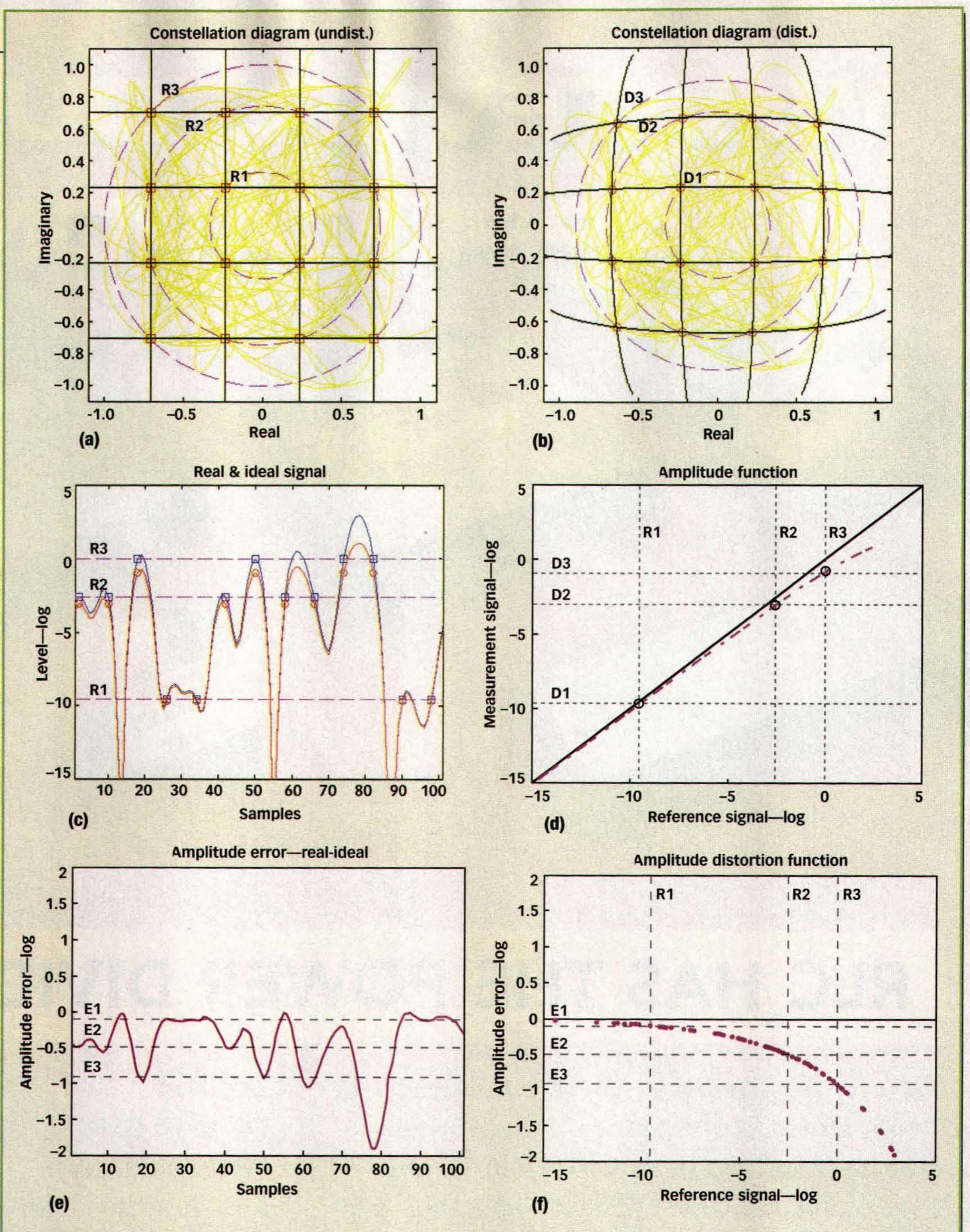
During analyzer operation, the ideal signal is reconstructed from the demodulated bits, so no previous knowledge of the transmitted data sequence or the ideal I/Q samples is necessary. The characteristics are determined according to the described scheme by comparing the ideal signal with the measurement signal. As a consequence, the amplifier will be measured in precisely the mode in which it is operated later on.

To compute the modulation error, the analyzer fits the measurement signal in a way that minimizes the RMS of the EVM at symbol times. This type of fitting is prescribed in the common mobile radio standards (e.g., EDGE[4]).

Figure 4(a) shows the error signal after the fitting, with the symbol times marked. In the logarithmic representation versus the reference signal, fitting causes a slight vertical shift of the measurement points and the interpolated compression curve [**Figs. 3(f) and 4(b)**].

After interpolation, the compression point is determined using two markers with a fixed horizontal distance of 10 dB between them. The point at which the vertical difference of the markers is 1 dB is determined by shifting the markers on the characteristic curve. The position of the marker (C) then represents the 1-dB compression point [Fig. 4(b)].

Figures 4(c) and 4(d) show a practical measurement of a 16-QAM modulation scheme with raised-cosine transmit filtering. This transmit filtering does not expect a receive filter and automatically results in intersymbol-interference-free (i.e., concentrated) constellation points. Fitting produces the following diagram: the position of the inside constellation points is moved slightly to higher levels. Constellation points with an average level are scaled virtually correctly, while the outside points with a high level are moved slightly inward.



3. These plots show (a) a 16QAM constellation diagram for an ideal signal (the three amplitude stages at the symbol time are marked by the circuits with radii R_1 to R_3, (b) a constellation diagram of output signals from an actual amplifier, (c) levels for real and ideal signals (with the amplitudes of the undistorted signals at the symbols times markets by the horizontal lines, R_1, R_2, and R_3), (d) the amplitude characteristics as an x/y representation of the two time signals (amplitudes at the symbol times are mapped to the R_1/D_1, R_2/D_2, and R_3/D_3 intersections), (e) the level difference over time, and (f) the AM/FM distortion characteristics as an x/y representation of the level drop versus the amplitude of the ideal signal (with the symbol times mapped to the R_1/E_1, R_2/E_2, and R_3/E_3 intersections).

The AM/AM distortion curve of the amplifier is obtained by interpolating all the measurement points [top half of Fig. 4(d)]. The bottom diagram of Fig. 4(d) shows the AM/PM curve, which is interpolated from the x/y representation of the phase difference versus the level of the ideal signal. Both characteristics are vertically shifted as a result of the fitting, but the differential computation of the compression point always provides the correct numeric value.

This new distortion measurement method can be used with all linear modulation schemes and any type of transmit filters. However, this method requires a measurement signal that is not receive filtered. Any receive filtering with strong band limitation causes the nonlinear effects to be distributed with the filter-impulse response over a number of

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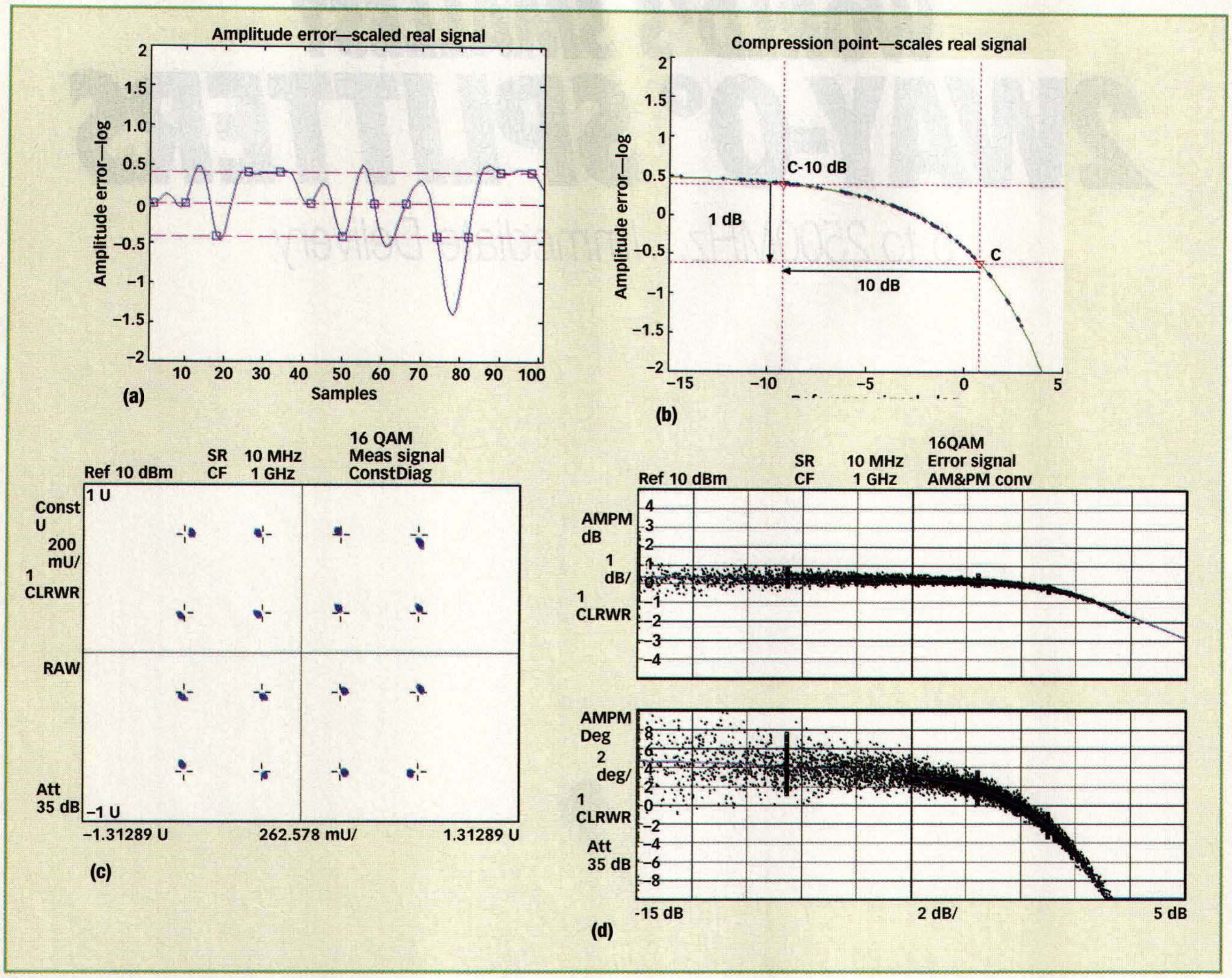

4. These plots show (a) the level different over time after fitting, (b) how the 1-dB compression point is determined from the interpolated characteristic curve, (c) an example of a distorted 16QAM constellation from the test amplifier's output signals, and (d) the AM/AM and AM/PM distortion characteristics.

symbol periods. As a result, the signal characteristic will be corrupted.

To illustrate the new distortion measurement method, a burst signal based on the EDGE mobile radio standard was used as an example. The digital standard EDGE uses a 3π/8-8PSK modulation scheme. For the transmitter,[4] a special filter is defined, which is not intersymbol-interference-free. As part of the example measurement, EDGE bursts were demodulated and the result ranges were aligned according to the position of the synchronization sequence (midamble) and limited to the valid area within the burst (useful part). Thus, the edges and areas outside the burst were not used for the measurement analysis.

For measurements performed on a wideband, bipolar small-signal amplifier (not shown), the vector signal analyzer computed the applied input power from the samples, determined the compression point and phase error and displayed them in absolute scaling. For this amplifier, the computed 1-dB compression point was found to be +10.36 dBm (DUT output level) with a phase distortion of 8.71 deg. Besides these level and phase characteristics, the comparison of the mean power levels and the crest factors (peak to average power) provides further information regarding the distortion behavior of the DUT. These measurements indicate a mean power compression of 0.68 dB and a decrease of the crest factor by 0.82 dB.

State-of-the-art vector signal analyzers make it easy to measure nonlinear distortion characteristics and modulation-dependent compression parameters. The same test setup can be used both for classic vector analysis and distortion measurements. The effectiveness of active predistortion for power amplifiers can be verified directly and not merely deduced from other test parameters such as EVM. MRF

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Design Finite Impulse Response Digital Filters

Precise FIR digital filters can be designed with the help of a popular mathematical modeling program and an understanding of basic performance trade-offs.

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Development Engineer
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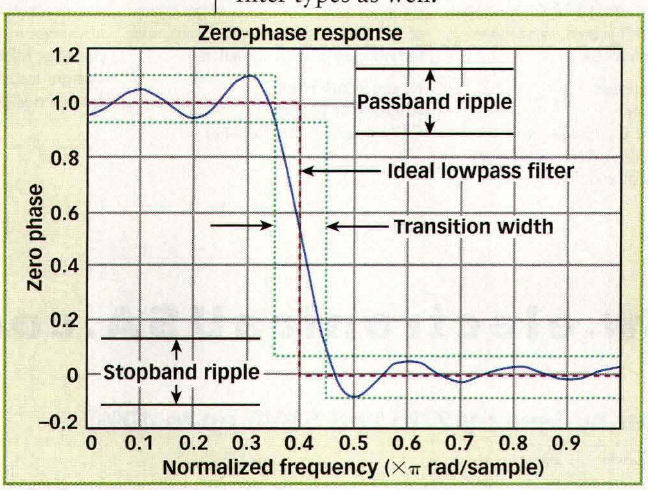
Digital filters have enabled much of modern communications and the measurements that support them. While most high-frequency design engineers may be familiar with the use of the MATLAB mathematical modeling tool from The MathWorks (Natick, MA) for the design of advanced antennas, the software also features powerful algorithms and toolboxes for the design of finite-impulse-response (FIR) digital filters. This four-part article series on designing FIR digital filters with MATLAB will open with a review of key FIR filter specifications and examples of optimal linear- and nonlinear-phase FIR filter designs. Although many lowpass filters will be shown, the design techniques apply to other filter types as well.



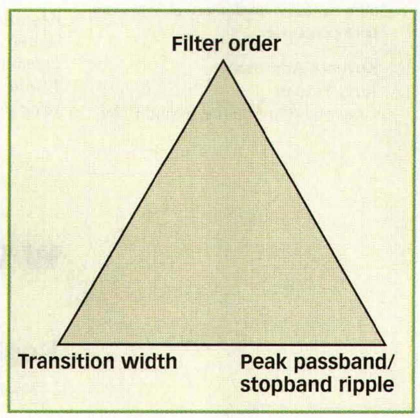
1. This illustration shows the typical deviations from the ideal lowpass filter when approximating the response with an FIR filter, $\omega_c = 0.4\pi$.

An ideal lowpass filter passes all frequency components of a signal below a designated cutoff frequency, ω_c, and rejects all frequency components of a signal above ω_c, according to:

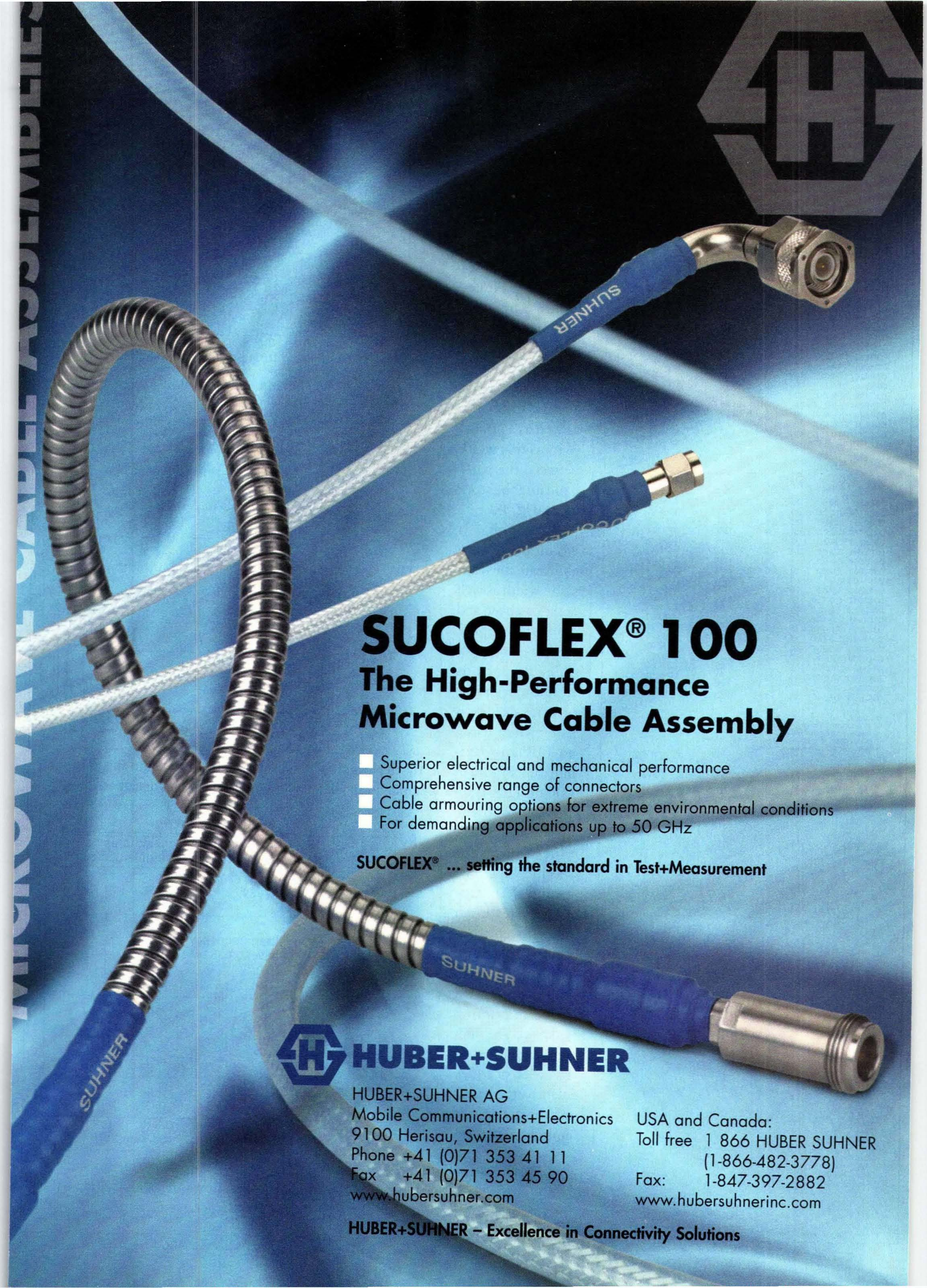
$$H_{LP}\left(e^{j\omega}\right) = \begin{cases} 1, & 0 \le \omega \le \omega_c \\ 0, & \omega_c < \omega \le \pi \end{cases} \quad (1)$$

The impulse response of an ideal lowpass filter in Eq. 1 can be found from[1]:

$$h_{LP}[n] = \frac{\sin(\omega_c n)}{\pi n}, \quad -\infty < n < \infty \quad (2)$$



2. Trade-offs among key FIR filter design specifications can be represented in the form of a triangle.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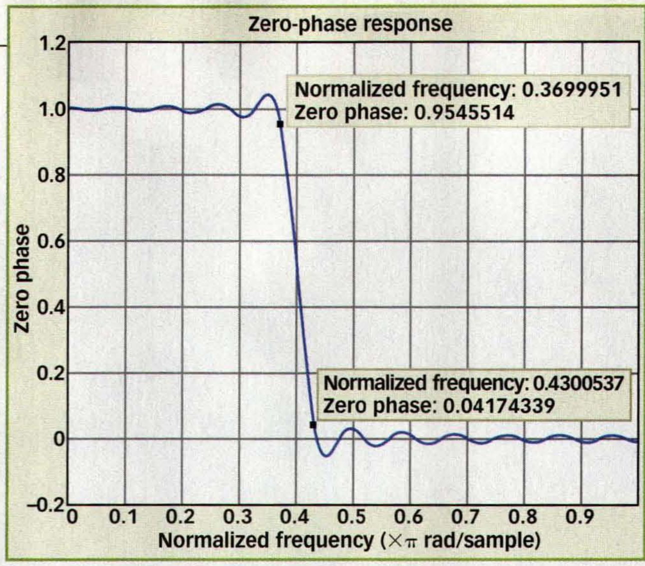
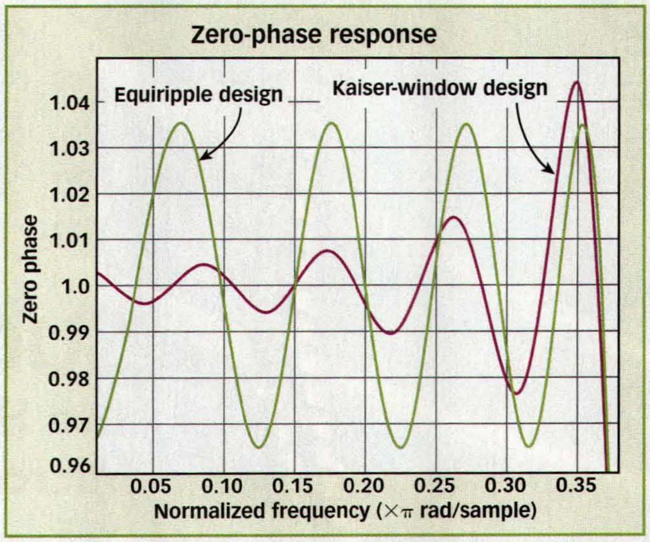
Because the impulse response required to implement the ideal lowpass filter is infinitely long, it is not possible to design an ideal FIR lowpass filter. Finite-length approximations of the ideal impulse response lead to ripples in both the passband and the stopband, as well as a nonzero transition width between the passband and the stopband **(Fig. 1)**.

Although it would be ideal to eliminate the transition width as well as the passband and stopband ripples, they are unavoidable deviations from the response of an ideal lowpass filter when using an FIR approximation. Practical FIR designs typically consist of filters meeting certain design specifications. For example, a practical FIR filter design has a given transition width and passband and stopband ripples set at some maximum value(s). The filter is also defined by its order, or equivalently, the length of the truncated impulse response.

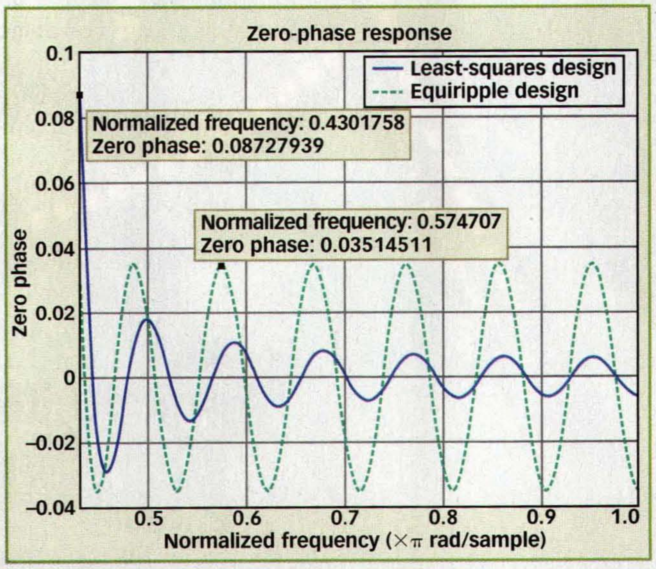
The trade-offs among key filter specifications can be represented in



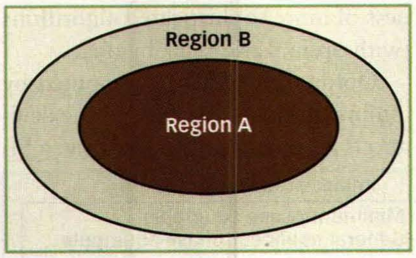
3. This response plot shows a Kaiser window design meeting predescribed specifications.



4. These plots show the passband ripple for both the Kaiser-window-designed FIR filter and the "remez"-designed filter.



5. This comparison of an optimal equiripple FIR design and an optimal least-squares FIR design shows that the equiripple filter has a smaller peak error but larger overall error.



6. The solution space for linear-phase and nonlinear-phase FIR filters for a given set of specifications can be represented by this diagram in which Region B represents the set of all linear- and nonlinear-phase FIR filters that meet the specifications.

the form of a triangle **(Fig. 2)**. The triangle shows the degrees of freedom available when designating design specifications. Because the sum of the angles is fixed, values can only be selected for two of the specifications; the third will be dictated by the selection of the other two. As with the triangle, if one of the specifications is made larger or smaller, it will impact one or both of the other two specifications.

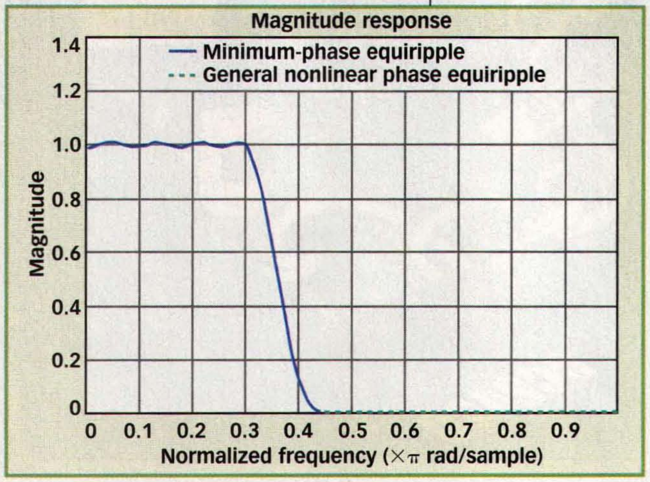
For example, consider the design of an FIR filter that meets the following specifications: a cutoff frequency of 0.4π rads/sample, a transition width of 0.06π rads/sample, and maximum passband and stopband ripple of 0.05 dB. The filter can easily be designed with the truncated-and-windowed impulse response algorithm implemented in the "firl" function within MATLAB's Filter Design Toolbox (or by using the "fdatool" graphical user interface) provided that a Kaiser-window filter design is used. **Figure 3** shows the zero-phase response of a filter designed according to these specifications. Since the transition width and peak-ripple performance are fixed, the filter order has been determined by these other two parameters.

Examination at the stopband-edge frequency $\omega_s = 0.43\pi$ shows that the peak passband/stopband ripples are within the allowable specifications. Usually, the specifications are exceeded because the order is rounded to the next integer greater than the actual value required.

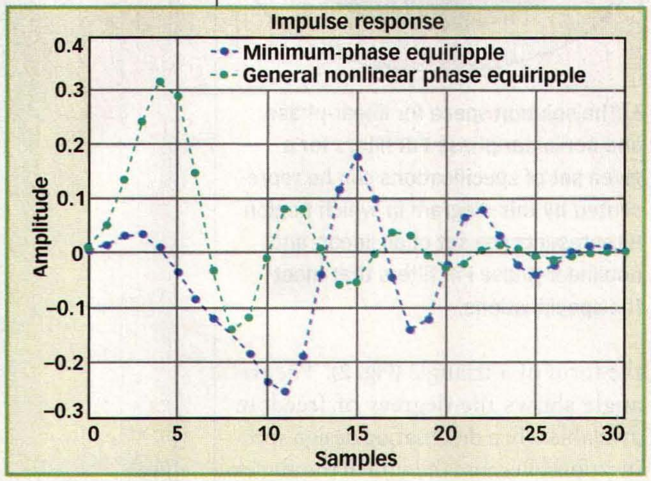
While the truncated-and-windowed

impulse response design algorithm is simple and reliable, it is not optimal. It yields designs that are generally inferior to those produced by algorithms employing optimization criteria; the nonoptimized designs will have greater order, greater transition width, or greater passband/stopband ripples than an optimized design. Since improved performance is usually desirable, the usefulness of more sophisticated algorithms (with optimization) is obvious.

Optimal designs are computed by minimizing some measure of the deviations between an ideal filter and a fil-



7. These magnitude responses compare a minimum-phase equiripple 30th-order filter and a general nonlinear-phase equiripple filter of the same order both designed to meet the same specifications.



8. These plots compare the impulse responses of an equiripple minimum-phase filter and a nonlinear-phase equiripple filter with virtually the same magnitude response.

ter to be designed. The most common optimal FIR design algorithms are based on fixing the transition width and the order of the filter, and minimizing the ideal/practical deviations of the passband/stopband ripples. This deviation or error can be expressed mathematically as (ref. 2):

$$E(\omega) = H_a(\omega) - H_{LP}(e^{j\omega}),\ \omega \in \Omega \qquad (A)$$

where:

$H_a(\omega)$ = the zero-phase response of the designed filter and

$$\Omega = [0, \omega p] \cup [\omega s, \pi]$$

It is still necessary to define a measurement to determine "the size" of $E(\omega)$, the quantity to be minimized as a result of the optimization. The most often used measures are $L\infty$-norm ($\|E(w)\|_\infty$ minimum-maximum designs) and L_2-norm ($\|E(w)\|_2$ least-squares designs).

To allow for different peak ripples in the passband and stopband, a weighting function $[W(\omega)]$ is usually introduced:

$$E_W(\omega) = W(\omega)\left[H_a(\omega) - H_{LP}(e^{j\omega})\right],\ \omega \in \Omega \quad (B)$$

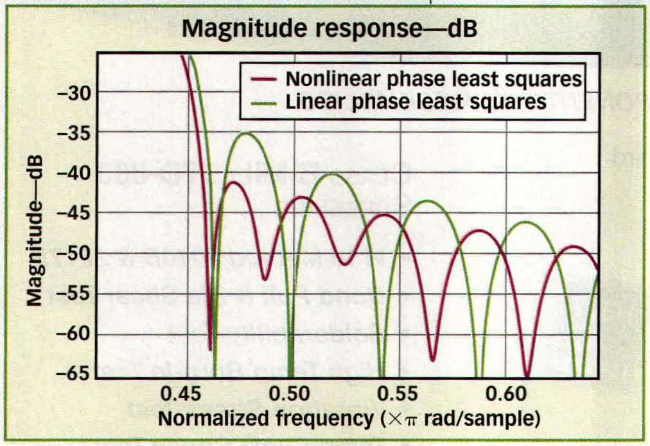
A filter with linear-phase response is desirable in many applications, notably image processing and data transmission. One of the desirable characteristics of FIR filters is that they can be designed very easily to have linear phase. It is well known[3] that linear-phase FIR filters will have impulse responses that are either symmetric or antisymmetric. For these types of filters, the zero-phase response can be determined analytically,[3] and the filter design problem becomes a well-behaved approximation theory problem.[4] The approximation involves determining the best solution for a given function (the response of an ideal lowpass filter) by means of a polynomial (the FIR filter) of given order. The "best" solution is one that minimizes the differences for a given performance parameter between an ideal filter and the filter to be designed.

The "remez" function in MATLAB

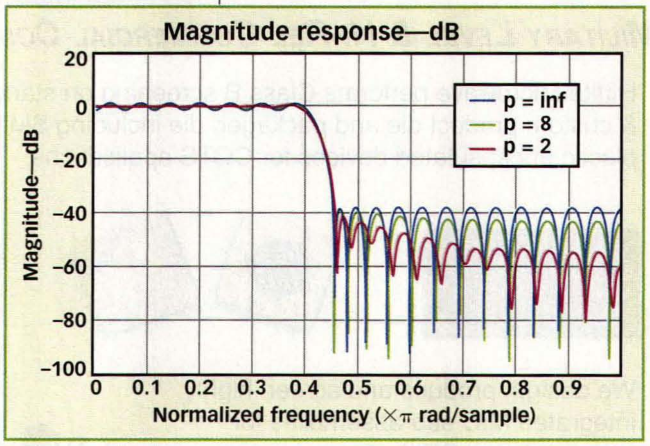
implements an algorithm developed in ref. 5 that computes a solution to the design problem for linear-phase filters in the L∞-norm case. The design problem is essentially to find a filter that minimizes the maximum error between the ideal and the actual filters. This type of design leads to so-called equiripple filters, i.e., filters in which the peak deviations from the ideal response are all equal.

The software's "firls" function implements an algorithm to compute solution for linear-phase FIR filters in the L_2-norm case. The design problem is to find a filter that minimizes the energy of the error between ideal and actual filters.

Linear-phase equiripple filters are desirable because they have the smallest maximum deviation from the ideal



9. This stopband comparison of a nonlinear-phase least-squares filter and a linear-phase least-squares filter of the same order show that the nonlinear-phase filter provides a smaller transition width and larger stopband attenuation.



10. These plots show optimal L_p-norm designs for different values of p. All filters have the same order and transition width.

filter when compared to all other linear-phase FIR filters of the same order. Equiripple filters are ideally suited for applications in which a specific tolerance must be met, such as when designing a filter with a given minimum stopband attenuation or a given maximum passband ripple.

For example, the earlier Kaiser-window design example was of 42nd order. With this same order, an equiripple filter (with fixed transition width) can be designed that is superior to the Kaiser-window design:

br = remez(42,[0 0.37 0.43 1], [1 1 0 0]);

Figure 4 shows the superposition of the passband details for the filters designed with the Kaiser window and with the "remez" function, with the maximum deviation appearing smaller for the "remez" design. In fact, since the filter is designed to minimize the maximum ripple (minimax design), the optimization process guarantees that no other linear-phase FIR filter of the 42nd order will have a smaller peak ripple for the same transition width.

Equiripple designs may not be the best design approach to minimize the error between the ideal and the designed filter's passband and stopband energy. A least-squares design approach is preferable to reduce the energy of a signal as much as possible in a certain frequency band. For example, for the same transition width and filter order as the earlier equiripple filter, a least-squares FIR design can be computed from:

bls = firls (42, [0 0.37 0.43 1], [1 1 0 0]);

The stopband energy for this case is:

$$E_{sb} = \frac{1}{2\pi}\int_{0.43}^{\pi} \left|H_a\left(e^{j\omega}\right)\right|^2 d\omega \qquad (C)$$

where:

$H_a(e^{jv})$ is the frequency response of the filter.

In this case, the stopband energy for the equiripple filter is approximately 1.7608e-004 while the stopband energy for the least-squares filter is 3.3106e-005. (As a reference, the stopband energy for the Kaiser-window design for this order and transition width is 6.1646e-005.) While the equiripple design has less peak error, it has more total error, measured in terms of its energy. The stopband details for both the equiripple design and the least-squares design are shown in **Fig. 5**.

An advantage of an FIR filter compared to an infinite-impulse-response (IIR) filter is the ability to attain exact linear phase in a straightforward manner. As mentioned previously, a linear-phase characteristic implies a symmetry or antisymmetry property for the filter coefficients. Nevertheless, the symmetry of the coef-

ficients constrains the possible designs that can be attained. For a filter with N + 1 coefficients, only N/2 + 1 of these coefficients is freely assignable (assuming that N is even). The remaining N/2 coefficients are immediately determined by the linear-phase constraint.

This can be thought of as reducing the search space for an optimal solution. The idea is depicted in **Fig. 6.**[7] Region A of the graph represents the set of all linear-phase FIR filters that meet a given set of specifications. This set contains both the optimal equiripple and the optimal least-squares filters mentioned so far. Region B represents the set of all FIR filters that meet a set of specifications, regardless of their phase characteristic, with Region A contained within Region B.

Without the linear-phase constraint (i.e., if the application at hand does not require a linear-phase characteristic), it is possible to design minimum-phase equiripple filters that are superior to optimal equiripple linear-phase designs based on a technique described in ref. 8. For example, the following minimum-phase design has both smaller peak passband ripple and smaller peak stopband ripple than the linear-phase equiripple design:

bm = gremez (42, [0 0.37 0.43 1], [1 1 0 0], [1,10], 'minphase')

It should be noted that this is not a totally unconstrained design. The minimum-phase requirement restricts the resulting filter to have all of its zeros on or inside the unit circle.[2]

Due to the requirement on the loci of the zeros, minimum-phase designs are not completely unconstrained. A general nonlinear-phase design algorithm is provided in the "firlpnorm" function. Consider the following specifications: a cutoff frequency of 0.375π rads/sample, a transition width of 0.15π rads/sample, maximum passband ripple of 0.008, and maximum stopband ripple of 0.0009. A 30th-order FIR equiripple filter (with nonlinear phase) can be designed to meet this set of specifications using the "firlpnorm" function:

blp = firlpnorm (30, [0 .3 .45 1], [0 .3 .45 1], [1 1 0 0], [1 1 10 10]);

This 30th-order design is in sharp contrast to a linear-phase filter, which would require a 37th-order architecture. By comparison, a minimum-phase equiripple filter designed using the "gremez" function would also require a 30th-order filter to meet the specifications.

The fact that two different nonlinear-phase filters of the same order meet the same specifications illustrates the difficulty associated with nonlinear-phase designs in general. There is no longer a unique opti-

mal solution to a given design problem. **Figure 7** shows the virtually identical magnitude responses. In contrast, **Fig. 8** shows the different impulse responses.

The "firlpnorm" function also provides the ability to select a different norm for the optimization. While the default optimization is for the L_∞-norm design, any design between (and including) L_2-norm and L_∞-norm is possible.

By the arguments given above, it is possible to attain a superior design using the "firlpnorm" function instead of the "firls" function for the same filter order, provided linear phase is not a requirement. For example,

b = firlpnorm (40, [0 .4 .45 1], {0 .4 .45 1], [1 1 0 0], [1 1 10 10], [2,2]);

b2 = firls (40, [0 .4 .45 1], [1 1 0 0], [1 20]);

yields a smaller transition width and a larger stopband attenuation for the nonlinear-phase case (with approximately the same peak passband ripple). The stopband details are shown in **Fig. 9**.

Because it is possible to choose the L_p-norm approach with which to optimize, the "firlpnorm" function is very flexible and allows for the designer to reach a compromise between equiripple and least-squares designs **(Fig. 10)**.

Because of the symmetry in the coefficients, some practical implementations will allow for a linear-phase response using roughly one-half the number of multipliers. This is particularly true with field-programmable gate arrays (FPGAs) and specialized hardware. The end result is that it may be possible to stick to a linear-phase design and achieve a more efficient implementation than comparable nonlinear-phase designs.

Next month, this four-part FIR filter design series will continue with an examination of optimal equiripple FIR filter designs, notably those in which both the peak ripples and the transition width are fixed. Part two will also explore the use of advanced filter design algorithms for implemented interpolated FIR filters. MRF

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Evaluate The Performance Of Amplifying Predistorters

The overcompensated feedforward (OCFF) approach to amplifier linearization can help save size and cost and improve the overall efficiency of linear power-amplifier systems.

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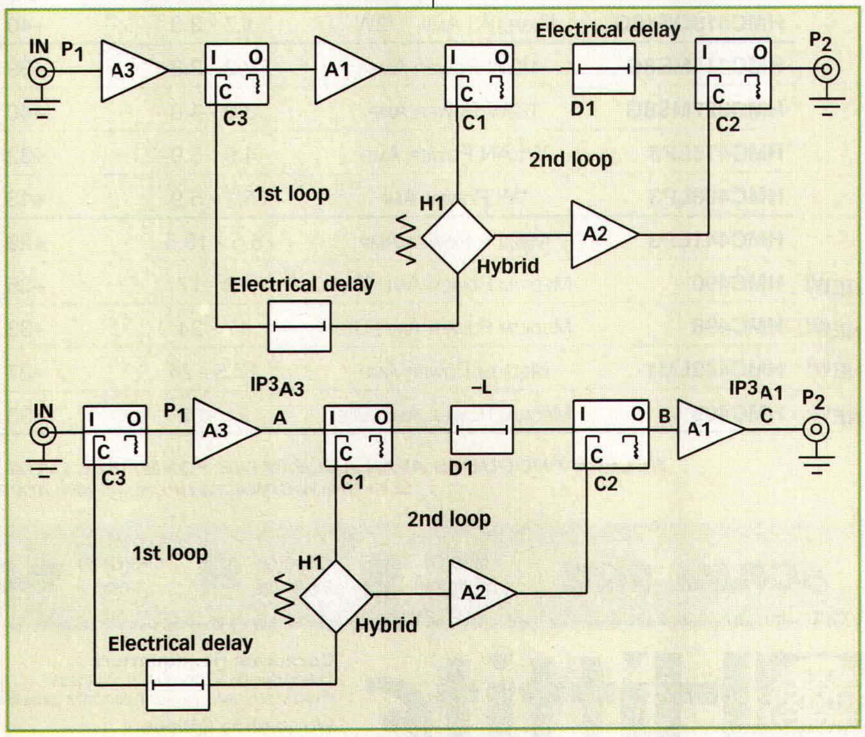
Predistortion circuits can improve the linearity of power amplifiers (PAs) and other devices used in communications systems. An amplifying predistorter, for example, employs the nonlinear distortion components of a preamplifier to correct the distortion of a following nonlinear device, such as a PA, rather than using separate nonlinear elements such as diodes or field-effect transistors (FETs). Amplifying predistortion, which can be considered a generalization of feedforward (FF) technique, can be used to improve amplifier efficiency while reducing cost and size compared to traditional FF methods. The approach can be adapted to systems with frequency upconverters



1. These block diagrams shows (a) a conventional feedforward system and (b) the overcompensated feedforward (OCFF) system.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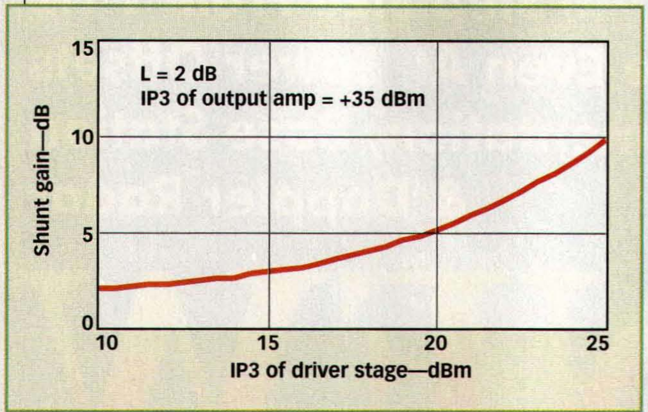


2. The required gain for the shunt path is plotted here as a function of driver-stage third-order intercept point.

as well as to systems with broadband electro-optic devices.

FF systems date from the early 20th century,[1,2] but their application in high-frequency systems began with the work of Seidel *et al.*[3] Since then, numerous FF systems have been reported in the literature for modern applications.[4-6] An example of amplifying predistortion (APD) can be found in ref. 7, where a driver amplifier is used to generate distortion to correct the distortion of the main PA. In this example, discussed in relationship to Class AB devices and operating over a narrow bandwidth, modest (9 dB) suppression of intermodulation distortion (IMD) was achieved.

APD is an attractive for several reasons. It does not suffer from the bandwidth restriction of a typical digital predistorter. In fact, it has been successfully applied to linearize devices spanning more than an octave. In addition, the APD approach has the following advantages over a conventional FF system:

1. Losses due to the delay line and couplers at the output are eliminated.
2. The error amplifier is of substantially lower power than used with a traditional FF system.
3. Since delay elements are not at the output, they need not exhibit low loss; compact, low-cost delay lines can therefore be used.
4. The method is amenable to adaptive linearization to correct for aging, drift, and temperature effects.
5. It is possible to linearize devices with inputs having a different nature than the outputs. Examples include a millimeter-wave amplifier preceded by a frequency upconverter and electro-optical transducers such as lasers or electro-optic modulators.

For practical application of APD, it would be useful to estimate the amount of IMD suppression that can be achieved by this technique compared to high-performance FF techniques. By treating the APD approach as a generalized form of FF, it may be possible to arrive at some estimates. **Figure 1(b)** shows a basic block diagram for the APD scheme; it bears a superficial resemblance to the FF system shown in **Fig. 1(a)**.

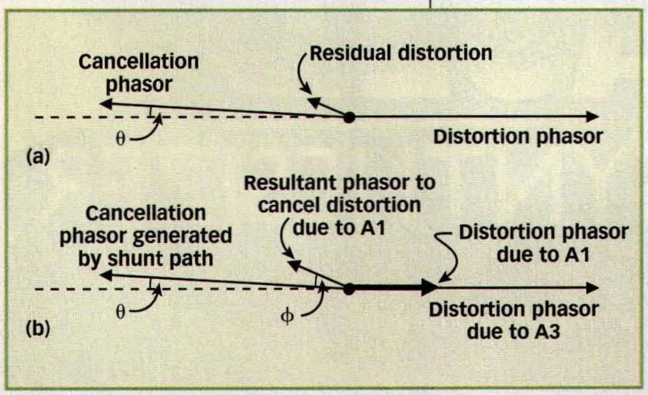
However, unlike an FF system where the distortion generated by a given device is canceled at its output, in an APD a second device (A3, the driver stage) is used as an additional source for distortion components. The distortion from A3 is fed with the correct magnitude and phase at the *input* of power stage A1 in order to cancel the distortion components generated by A1. If A1 is removed from the circuit, the distortion from the path consisting of A2 does not completely cancel the distortion from the main path as in a FF scheme. But the extra gain of A2 helps generate more distortion than necessary compared to a FF scheme, and this extra distortion cancels the distortion from the power stage (A1). For this reason, this technique can be called the over-compensated feedforward (OCFF) technique.

In order to develop a simplified OCFF analysis approach, fifth- and higher-order distortion components will be neglected, a reasonable assumption for Class A amplifiers and certain electro-optic components. It will also be assumed that the imaginary part of the devices' nonlinear transfer function is negligible.[8,9]

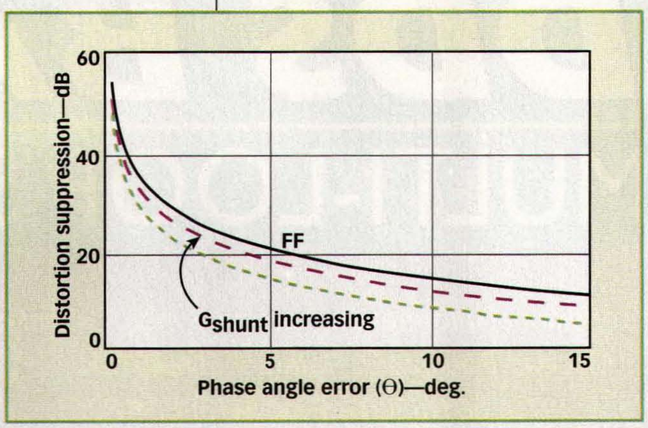
Referring to Fig.1(b), the following parameters can be defined as:

P_1 = the input power (two-tone excitation) at the input port of A3 (in dBm/tone);

$IM3_{A1}$ and $IM3_{A3}$ = the two-tone third-order intermodulation components at the outputs of A1 and A3, respectively (in dBm);



3. These phasor diagrams shows (a) the feedforward scheme and (b) the OCFF scheme. All phasors are referred to point B of Fig. 1(b).



4. Distortion suppression is shown here as a function of the phase angle error.

$IP3_{A1}$ and $IP3_{A3}$ = the output third-order intercepts of A1 and A3, respectively (in dBm);

G_{A1}, G_{A2}, and G_{A3} = the gain of A1, A2, and A3, respectively (in dB);

L = the loss of the main path, which is comprised of the insertion losses of couplers C1 and C2 and the loss through delay line D1 (in dB).

At point A of Fig.1(b), following the definition of third-order intercept point:

$$IM3_{A3} = 3(P_I + G_{A3}) - 2\,IP3_{A3} \qquad (1)$$

At point B, the intermodulation voltage signal from the main path is:

$$V_1 = 10^{\frac{IM3_{A3}-L}{20}} \qquad (A)$$

and the intermodulation voltage signal from the shunt path at point B is:

$$V_2 = 10^{\frac{IM3_{A3}+G_{shunt}}{20}} \qquad (B)$$

where:

G_{shunt} = the gain in the shunt path comprised of G_{A2} minus coupling losses in C1 and C2 and insertion loss in hybrid coupler H1.

If the delays in the main and shunt paths are identical, the phasors V_1 and V_2 would be in opposite phase, provided that gain G_{A2} exceeds some minimum value. In that case, the resultant voltage at point B can be expressed as:

$$V_1 - V_2 = 10^{\frac{3P_I}{20}} \times 10^{\frac{3G_{A3}-2IP3_{A3}}{20}} \times \left(10^{\frac{-L}{20}} - 10^{\frac{G_{shunt}}{20}}\right) \qquad (2)$$

At point C, the intermodulation voltage signal from A1 alone can be expressed as:

$$IM3_{A1} = 3(P_I + G_{A3} - L + G_{A1}) - 2IP3_{A1} \qquad (3)$$

Referring the above signal to the point B results in:

$$IM3_{A1} - G_{A1} = 3P_I + 3G_{A3} - 3L + 2G_{A1} - 2IP3_{A1} \qquad (C)$$

which can be converted to voltage to obtain:

$$V_3 = 10^{\frac{3P_I+3G_{A3}-3L+2G_{A1}-2IP3_{A1}}{20}} \qquad (4)$$

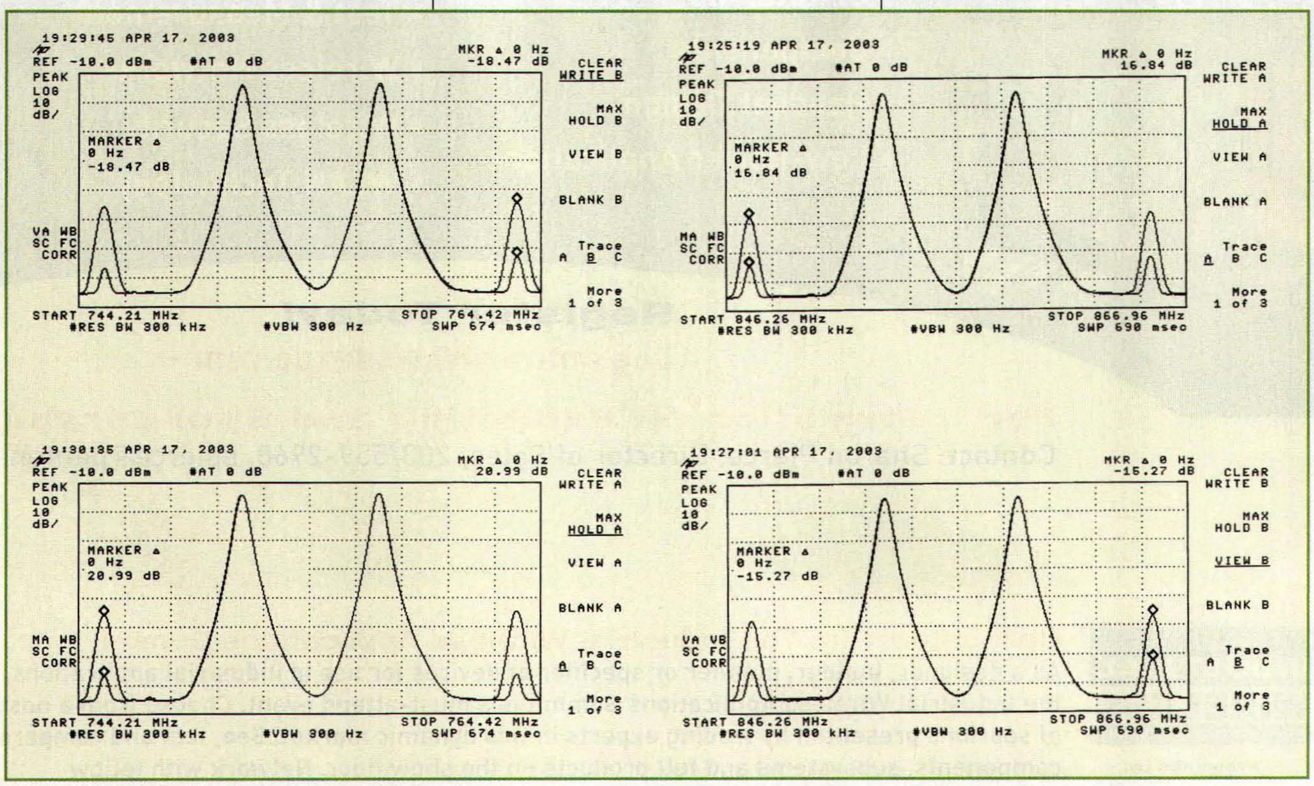
Cancellation of the distortion phasors requires that:

$$V_1 - V_2 + V_3 = 0 \qquad (5)$$

From Eqs. 2, 4, and 5, it is possible to solve for G_{shunt} to get:

$$G_{shunt} - L + 20\log\left(1 + 10^{\frac{(IP3_{A3}-L)-(IP3_{A1}-G_{A1})}{10}}\right) \qquad (6)$$

It can be seen that $IP3_{A3} - L$ is the third-order intercept at point B from amplifier A1 and $IP3_{A1} - G_{A1}$ is the third-order intercept from amplifier A1, also referenced to point B. If $IP3_{A3} - L = IP3_{A1} - G_{A1}$, then $G_{shunt} = -L + 6$ which can be interpreted as follows. The extra gain in the shunt path overcomes the loss L (as in FF) and further doubles the distortion signal to compensate the distortion from A1. For the hypothetical case of A1 being perfectly linear ($IP3_{A1}$ going to infinity), the OCFF scheme degenerates to a traditional FF scheme, and G_{shint} becomes –L. **Figure 2** graphically depicts shunt gain as a function of the driver-stage out-



5. These plots show the distortion suppression performance of the predistorter at (a) 700 MHz and at (b) 800 MHz.

put third-order intercept point.

An estimate of the figure of merit of the OCFF scheme can be made based on the assumption that the distortion phasors are perfectly matched in magnitude but possesses a phase error, θ, between them. **Figure 3(a)** shows the phasor diagram of a FF scheme, where the suppression of the distortion phasor with respect to the original distortion phasor is calculated from elementary geometry to be $20\log[\sin(\theta/2)]$. **Figure 3(b)** shows the phasor diagram for the OCFF scheme, with all phasors referred to point B of Fig. 1(b). The resultant phasor due to the distortion from A1 and the shunt path is at an angle ϕ from the reference as shown. Assuming that the phase angle of the distortion phasor from A1 is zero, the suppression of the final distortion phasor with respect to the original distortion phasor is given by $20\log[\sin(\phi/2)]$. From simple geometry, it follows that:

$$Suppression_{OCFF} = 20\log \left[\sin\frac{1}{2}\left(\arctan\frac{G_{shunt}\times\sin\theta}{G_{shunt}\times\cos\theta - 1}\right)\right] \quad (7)$$

Figure 4 shows the suppression as function of phase angle error θ. It can be seen that OCFF is always inferior to FF as far as this figure of merit goes, approaching FF performance for infinitely large values of G_{shunt}. However, large G_{shunt} values also imply low levels of distortion from A1, thereby defeating the purpose of the OCFF approach. The amount of noise injected into the system also becomes large for high values of G_{shunt}. Nevertheless, the absence of loss at the output of A1 in the OCFF approach more than compensates for this reduction in suppression.

If the imaginary part of the active devices' nonlinear transfer function is negligible, distortion cancellation on the order of 20 dB is realistic. Amplifier A2 must be linear, but does not have to deliver high amounts of power compared to the traditional FF scheme since it is feeding the distortion components to the input of the power stage rather than at its output.

For Class AB operation, fifth- and higher-order IMDs must be considered. Modeling may be further complicated when the phases of the IMDs become dependent on the power level. Therefore, the problem reduces to matching the nonlinear transfer function of devices A1 and A3 in an appropriate manner to achieve optimum suppression of the IMDs (a topic for future discussion). It should be noted that the method mentioned for Class A operation is a special case of this generalized approach. Intuitively, OCFF operation will be optimized when A1 and A3 will operate with the same conduction angles.

Assuming output power of P_2 in both cases, the efficiencies of the FF and OCFF techniques can be compared. The parameters of interest are pure numbers (not in units) and can be expressed as:

g_{A1}, g_{A2}, and g_{A3} = the gains of amplifiers A1, A2, and A3, respectively;

c1, c2,and c3 = the coupling coefficients of couplers C1, C2, and C3, respectively (less than 1); and

l_m = the loss in the main path, given by $l_m = (1 - c1)l_d(1 - c2)$

where:

l_d = the loss in the main path delay line.

Power dissipation in A3 plays a negligible role in computing the overall efficiency as long as gain g_{A1} is large enough. Therefore, only power dissipation in A1 and A2 will be considered. If η1 is the efficiency of amplifier A1, the DC power dissipation values by A1 for the FF and OCFF cases, respectively, are:

$$P_{dc1FF} = \frac{P_2}{l_m \times \eta_1} \quad (8a)$$

$$P_{dc1OCFF} = \frac{P_2}{\eta_1} \quad (8b)$$

To calculate the DC power dissipation in amplifier A2, two distinct cases will be considered. In case 1, distortion components at the output of H1 are orders of magnitude smaller than the distortion components of the main carriers, which is usually true for Class A operation. Therefore, A2 essentially handles the power from the suppressed carriers, determined by the amount of cancellation available from H1. The RF power levels at the output of A2 for the FF and OCFF cases, respectively, are P_2sg_{A2}/l_m and $P_2sg_{A2}/(g_{A1}l_m$, where s is the suppression on the carrier obtained at the output of H1.

Consequently, the DC power dissipation amounts by A2 for the FF and OCFF cases, respectively, are:

$$P_{dc2FF} = \frac{P_2 \times s \times g_{A2}}{l_m} \times \frac{1}{\eta_2} \quad (9a)$$

$$P_{dc2OCFF} = \frac{P_2 \times s \times g_{A2}}{g_{A1} \times l_m} \times \frac{1}{\eta_2} \quad (9b)$$

It should be noted that A2 is most likely to be a Class A amplifier since it must handle signals with large peak-to-RMS ratio. Therefore, although the amplifier for the FF case must be more powerful than that for the OCFF case, both would have same efficiency, η_2.

Designating η_{FF} and η_{OCFF} as the overall efficiencies for the FF and OCFF approaches, respectively, it is possible to determine from Eqs. 8(a) and 8(b) and 9(a) and 9(b) that:

$$\frac{\eta_{OCFF}}{\eta_{FF}} = \frac{1}{l_m} \times \frac{\frac{1}{\eta 1} + \frac{1}{\eta 2} \times s \times g_{A2}}{\frac{1}{\eta 1} + \frac{1}{\eta 2} \times \frac{s \times g_{A2}}{l_m \times g_{A1}}} \quad (10a)$$

Since $l_m g_{A1} > 1$, and $l_m < 1$, the efficiency of an OCFF system is always larger than that of an FF system.

In case 2 for the evaluation of efficiency, the distortion components at the output of H1 are large compared to the main carriers, which is often true for Class AB operation. For an FF system, the RF power delivered at the output of A2 is given by $P_2(\delta/c2)$, where δ is the ratio of the distortion power to the carrier power before linearization. For an OCFF system, the RF power at the output of amplifier A2 is $(P_2/g_{A1})(\delta/c2)o_f$, where o_f is the overcompensation factor (greater than 1) that accounts for the excess signal required as explained

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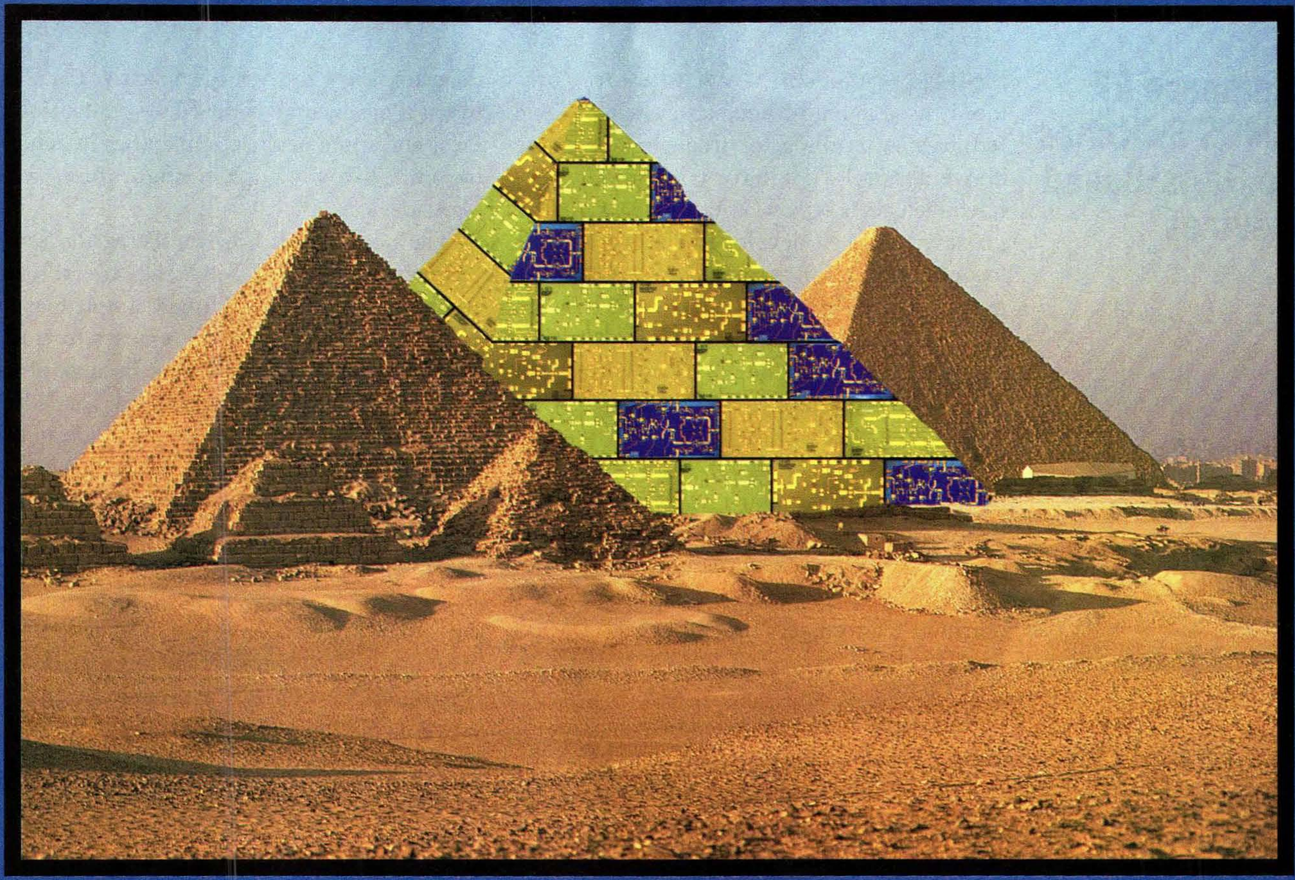

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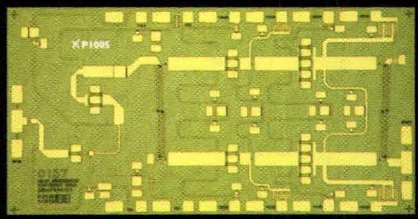

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Compare RF CMOS Switches To GaAs-Based Devices

HIGH-PERFORMANCE RF SWITCHES are commonly used in high-frequency systems. While requirements calling for frequency coverage above about 1 GHz have traditionally featured switches based on GaAs technology, a new process has been developed to leverage the low cost, repeatability, and high levels of integration offered by silicon CMOS, but with the high-frequency performance previously available only from GaAs devices. The Ultra-Thin Silicon-on-Sapphire (UTSi) CMOS process from Peregrine Semiconductor Corp. (San Diego, CA) supports high levels of analog and digital integration and the high-speed performance characteristics of GaAs substrates. It is detailed in a comparison of UTSi and GaAs switches contained in the five-page application note "Performance Benefits of UTSi CMOS RF Switch Technology Compared to GaAs Based RF Switches."

The application note opens with a review of GaAs transistor characteristics, including the complex power-supply requirements for on and off states often the need for off-chip circuit elements, such as coupling capacitors. Simple diagrams show the type of capacitive coupling that is typically used in the RF-to-control interface, and some of the complexities in achieving multithrow switch configurations with GaAs devices.

The note details how the Peregrine Semiconductor UTSi CMOS switches are fabricated and compares the relatively simple biasing and control-interface requirements to traditional monolithic GaAs designs. Because of the circuit integration possible with the UTSi CMOS process, it is a simple matter to include an on-chip micropower negative voltage generator to increase the operating range of devices fabricated on the CMOS process.

Copies of application note AN18, along with a variety of other application notes detailing some of the company's product lines, including phase-locked loops (PLLs), prescalers, and fractional-N frequency synthesizers, can be downloaded free of charge from the website.

Peregrine Semiconductor Corp., 6175 Nancy Ridge Dr., San Diego, CA 92121; (858) 455-0660, FAX: (858) 455-0770, Internet: www.peregrine-semi.com.

The note details how the UTSi CMOS switches are fabricated and compares the relatively simple biasing and control-interface requirements to traditional monolithic GaAs designs.

Aiding RF And DC Semiconductor Measurements

MEASUREMENTS OF RF AND DC device parameters at the semiconductor wafer level require large and complex test equipment. Fortunately, a new application note from Micromanipulator, "Measurements for Analog and Digital Circuits," discusses the challenges of assembling test setups for accurate on-wafer measurements.

The straightforward, four-page application note explains how vector-type current and voltage measurements can be used to calculate a variety of other related electrical characteristics, including signal gain and loss, reflections, reactance, timing, and phase shifts through a circuit. Unfortunately, there are no perfect or ideal test setups that always allow direct measurements on the wafer level, and usually test cables, adapters, and probes are needed to extract electrical signals and apply bias supplies to devices contained on a wafer. Because test cables have their own electrical responses, a test engineer must become familiar with a test setup, the frequency of operation, the probe or connection effects, and other contributing factors that can impact the overall measurement capabilities and accuracy of a test setup.

The application note describes how to make "bridged" measurements (and cautions about the limitations of these measurements), it discusses the errors caused by the inversion effects of quarter-wave lengths of "lossless" test cables, and makes recommendations for proper techniques for terminated measurements on a device under test (DUT). Although some of the company's probes are used as examples, the note is largely generic in nature and can be applied to equipment from any number of different measurement-equipment manufacturers.

Copies of the four-page application note are free upon request from the company by contacting Karen Schanhals at the toll-free number or by visiting the website.

Micromanipulator, 1555 Forrest Way, Carson City, NV 89706; (775) 882-2400, (800) 654-5659, FAX: (775) 882-7694, e-mail: info@micromanipulator.com, Internet: www.micromanipulator.com.

cover story

Load-Pull Tuners Are Frequency Selective

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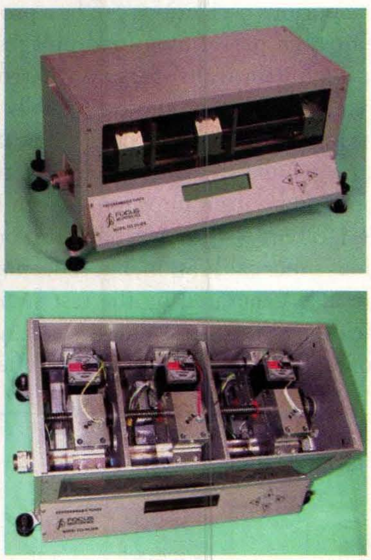
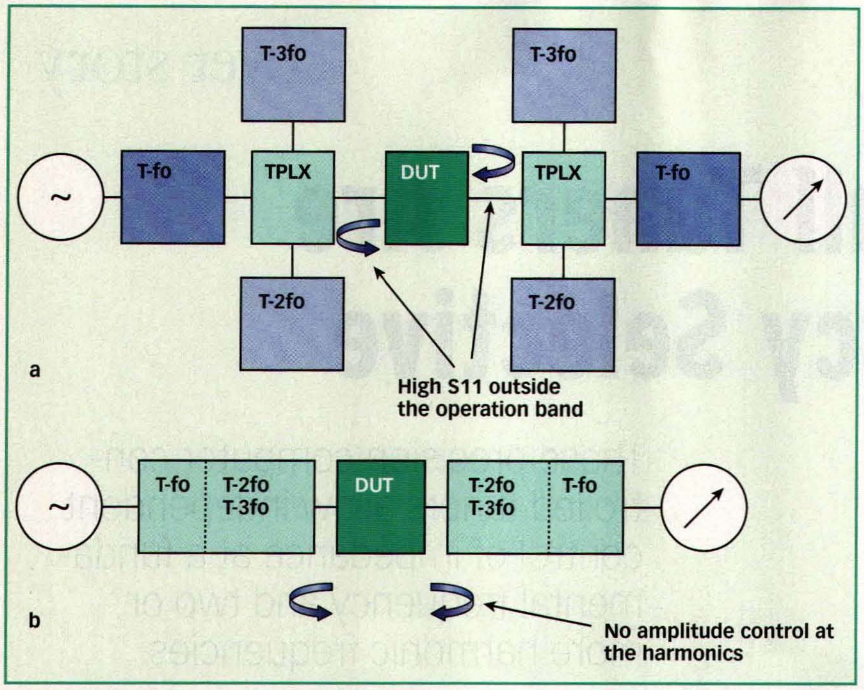
Power-amplifier (PA) designers must constantly strive for improved linearity and efficiency to meet the needs of current and emerging communications standards. High-data-rate digital modulation formats require nearly distortion-free performance from a PA, requiring the amplifier designer to fully characterize active devices before developing matching networks. Fortunately, a new series of frequency-selective tuners from Focus Microwaves, Inc. (Dollard-des-Ormeaux, Quebec, Canada) provide PA designers with the ability to create independently controllable impedances at three different frequencies for unprecedented insight into the nonlinear behavior of large-signal active devices.

The new frequency-selective tuners (FSTs) greatly simplify amplifier amplitude and phase tuning at fundamental and several harmonic frequencies **(Fig. 1)**. The tuners, which are compact enough for positioning with wafer-probing equipment, provide a high reflection coefficient at the reference plane of a device under test (DUT), whether in a test fixture or on wafer. The tuners feature low out-of-band reflections to minimize the change of DUT oscillation or spurious generation. Resonator cells are relatively transparent out of band and thus can be readily cascaded for true multifrequency harmonic tuning. The tuners are currently available for frequency ranges from 1.8 to 18.0 GHz.

Modern PA designers require the most accurate possible transistor data for a wide range of RF and DC parameters as functions of DC bias, RF impedance (at different frequencies), source power, and other variables. For example, power transistors are often driven into regions of strong nonlinearity in order to achieve high power-added efficiency (PAE). One way to reach the maximum PAE for given levels of linearity and output power involves using harmonic-load pull measurements to determine the impedance conditions at the fundamental frequency and the second-harmonic frequency that result in the best



1. The model FST-1818 is a frequency-selective tuner (a) designed for use from 1.8 to 18 GHz, with separate resonators (b) for independent impedance control at a fundamental and two harmonic frequencies.



2. A traditional harmonic load-pull system (a) can involve as many as six impedance tuners and supporting hardware, which can be simplified through the use of combination fundamental/harmonic tuners (b).

compromise among PAE, linearity, and output power. By optimizing the RF impedance at the second harmonic frequency, it is not uncommon to double the PAE compared to nonoptimized conditions.

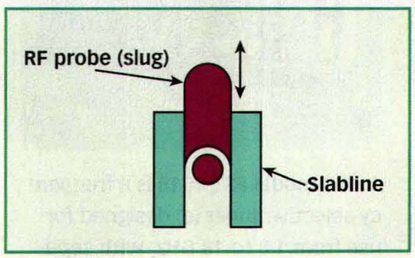
A full harmonic load-pull measurement system for testing at fundamental, second-harmonic, and third-harmonic frequencies incorporates DC sources, RF signal sources, power meters, driver amplifiers, passive signal-coupling components, analyzers, and six impedance tuners [**Fig. 2(a)**]. The complex setup is expensive and tedious to calibrate and operate, especially when running on-wafer measurements. A more practical system employs just two combination tuners capable of impedance tuning at the fundamental and harmonic frequencies [**Fig. 2(b)**], an approach developed several years ago by Focus Microwaves. Unfortunately, the concept of harmonic tuning used in these components only provides control of the phase of the reflection factor at harmonic frequencies, albeit at a very high reflection coefficient. For full amplitude control at all harmonic frequencies, a test set requires the use of triplexers or active loads, which yield a limited range of reflection coefficient, have inconveniently high reflection at lower frequencies (triplexers) with the associated risk of spurious oscillations with high-gain devices, or active loads, which are complex and expensive.[1]

Traditional microwave tuners employ a movable metallic and RF grounded probe inside a high-frequency slab line (**Fig. 3**). The distance between the probe and the center conductor of the line (the vertical position) determines the amplitude of the reflection coefficient seen at the tuner's ports. Moving the probe along the center conductor (the horizontal position) adjusts the phase. In general, such tuners provide almost constant VSWR over wide frequency ranges, such as 0.8 to 3.0 GHz and 3 to 18 GHz, with gradually roll off of VSWR at the high end of low reflection at frequencies below the specified operating range. Although wideband performance is desirable for generic test equipment, it often creates instability problems when performing variable impedance tests on high power/gain devices.

The frequency-selective tuners (FSTs) were developed to combine the advantages of frequency-selective frequency discriminators with lowpass tuners. The new tuners employ noncontacting resonant probes (slugs) at all frequencies (**Fig. 4**) and allow independent adjustment of reflection factors at a fundamental and two or three harmonic frequencies. Depending on the fundamental frequency, harmonics as high as the fourth or fifth harmonic can be selected. The limitation of two or three harmonic frequencies is due to the required travel of the probes (for 360-deg. phase coverage) and the allowable overall size of the tuner.

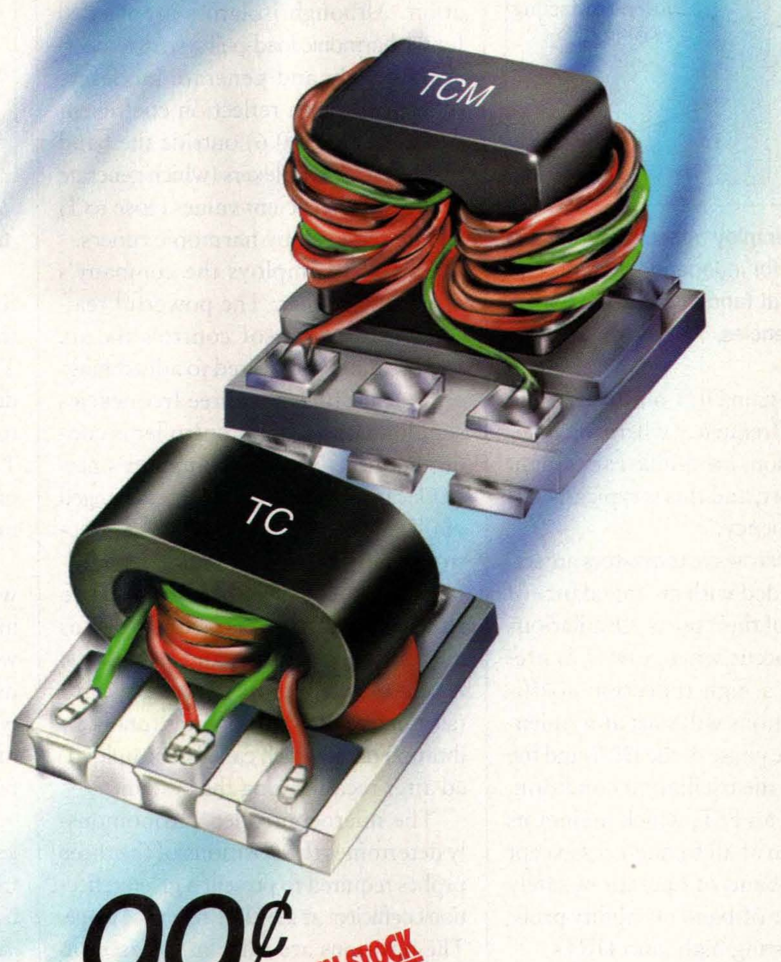
When the resonant probe is moved closer to the central conductor, coupling increases and so does the reflection factor. Due to the high quality factor (Q) of the resonator, the overall resonant frequency is not significantly affected (shifted), allowing a fairly well controlled tuning scheme for the fundamental frequency and second- and third-harmonic frequencies (**Fig. 5**). The tuner is calibrated at each frequency in the manner of an ordinary wideband tuner, resulting in well-behaved calibration patterns for amplitude and phase (**Fig. 6**). The company's load-pull software allows accurate interpolation and tuning to any point of the Smith Chart, and includes the capability of "back-tuning" to reduce the harmonic cross-tuning below –20 dB.

Spurious oscillations can be a problem when performing load-pull tests on high-gain transistors in a narrow frequency band of interest (usually at the low-frequency bandedge where the gain is highest). During testing, the input and output impedances seen by the DUT are well known, due to the previous tuner calibration, but reflections are also generated by the test setup, including the tuners, and these are not controllable. Because of these reflec-



3. A precision slide-screw tuner can provide a large range of reflection coefficients over a wide band of frequencies.

RF TRANSFORMERS



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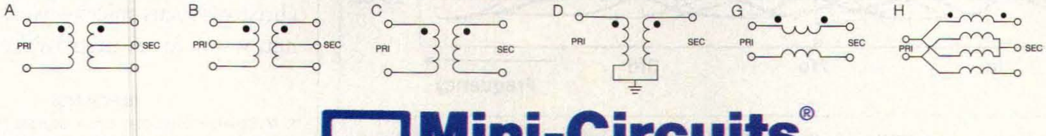
TC1-1	1C	1.5-500	5-350	1.19
TC1-15	1C	800-1500	800-1500	1.29
TC1.5-1	1.5D	.5-2200	2-1100	1.59
TC1-1-13M	1G	4.5-3000	4.5-1000	.99
TC2-1T	2A	3-300	3-300	1.29
TC3-1T	3A	5-300	5-300	1.29
TC4-1T	4A	.5-300	1.5-100	1.19
TC4-1W	4A	3-800	10-100	1.19
TC4-14	4A	200-1400	800-1100	1.29
TC8-1	8A	2-500	10-100	1.19
TC9-1	9A	2-200	5-40	1.29
TC16-1T	16A	20-300	50-150	1.59
TC4-11	50/12.5D	2-1100	5-700	1.59
TC9-1-75	75/8D	0.3-475	0.9-370	1.59

LEADS Plastic Base

(actual size) MODEL	Ω Ratio & Config.	Freq. (MHz)	Ins. Loss❖ 1dB (MHz)	Price $ea. (qty. 100)
TCM1-1	1C	1.5-500	5-350	.99
TCML1-11	1G	600-1100	700-1000	1.09
TCML1-19	1G	800-1900	900-1400	1.09
TCM2-1T	2A	3-300	3-300	1.09
TCM3-1T	3A	2-500	5-300	1.09
TTCM4-4	4B	0.5-400	5-100	1.29
TCM4-1W	4A	3-800	10-100	.99
TCM4-6T	4A	1.5-600	3-350	1.19
TCM4-14	4A	200-1400	800-1000	1.09
TCM4-19	4H	10-1900	30-700	1.09
TCM4-25	4H	500-2500	750-1200	1.09
TCM8-1	8A	2-500	10-100	.99
TCM9-1	9A	2-280	5-100	1.19

Dimensions (LxW): TC .15" x .15"   TCM .15" x .16"   ❖Referenced to midband loss.

ELECTRICAL CONFIGURATIONS



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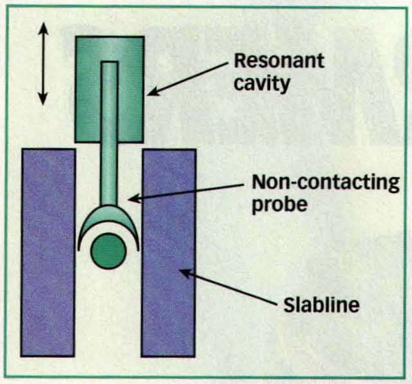


4. The FSTs employ noncontacting resonant probes for independent control of impedances at fundamental and harmonic frequencies.

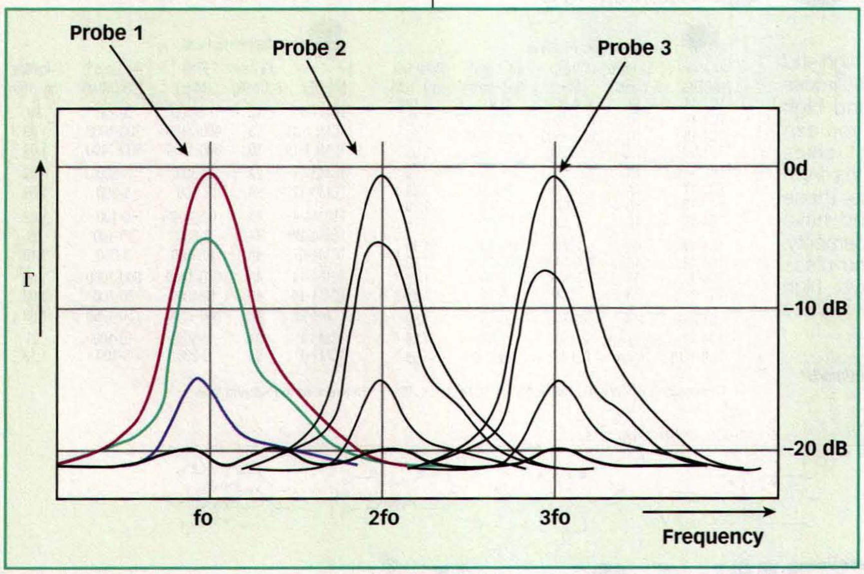
tions, a high-gain DUT may start oscillating at any frequency where the oscillation conditions are fulfilled at its input or output port, and this is typically not the test frequency.

Typical microwave transistors are stable when loaded with an impedance of 50 Ω at one of their ports. Oscillations most often occur when a DUT is presented with a high reflection coefficient; oscillations will start at frequencies where the phase of the DUT and the tuner satisfy the oscillation condition. Fortunately, an FST, which maintains low reflection at all frequencies except in a narrow band of operation, safely solves the out-of-band instability problem when testing high-gain DUTs.

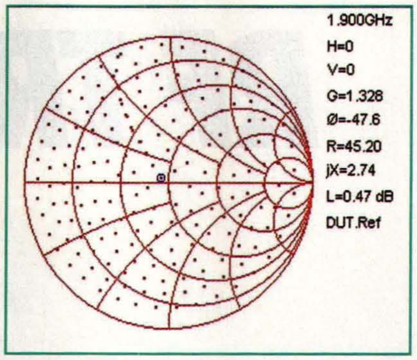
Passive components in the traditional harmonic load-pull measurement setup, such as narrowband isolators and low-loss triplexers, generate reflections outside their normal band of operation. Although isolators are needed for the harmonic load-pull system to operate properly and generate moderate reflections (with reflection coefficient values of 0.5 to 0.6) outside the band of interest, the triplexers (which generate reflection coefficient values close to 1) can be replaced by harmonic tuners.

The FSTs employs the company's iTuner controller. The powerful real-time microprocessor controls the six stepper motors required to adjust magnitude and phase at three frequencies simultaneously. The controller is connected directly to the computer's network card and commands are exchanged via TCP/IP. A nonvolatile 4K electronically erasable programmable read-only memory (EEPROM) is used to store operational parameters, such as model, serial number, and IP address. A removable data flash-memory card (as much as 64 MB) contains tuner calibration data, which can also be uploaded after recalibrating the system.

The microcontroller autonomously determines the positions of the three probes required to present a given reflection coefficient at the DUT reference plane. The positions are calculated based on calibration data for a given frequency, and take into account user-defined test-



5. The tuning scheme of an FST is designed to alter the amplitude without detuning the resonator frequency.



6. The plot shows FST amplitude/phase tuning at harmonic frequencies.

fixtures and cables inserted between the DUT and the tuner reference plane. The stepper motors are controlled via dedicated control ICs directly connected to the processor input/output (I/O) port. The acceleration/deceleration profile of the motors has been fine tuned to minimize vibration.

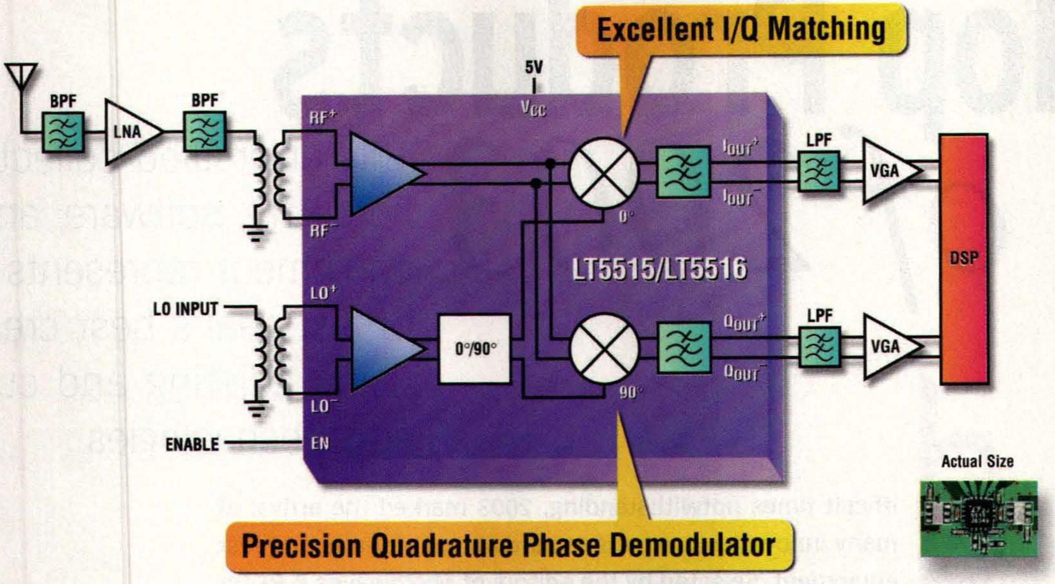
The tuners are factory calibrated, with calibration data are stored on internal flash memory. The system software allows re-calibration of the tuner using a calibrated vector network analyzer (VNA). A novel calibration technique coordinates a three-probe calibration resulting in a high-resolution 400 × 400 × 400 calibration data grid in less than 15 minutes for each fundamental frequency. The internal microprocessor uses the calibration data and second-order interpolation routines to calculate the reflection coefficient seen be the DUT for any valid probe position.

The FSTs operate via an enhanced version of the company's standard "WinPower" load-pull software for Windows-based personal computers. A toggle switch in the software's on-screen toolbox selects the resonator to tune. The corresponding frequency, axis positions, setup and tuner losses and actual impedance seen by the DUT are automatically switched and displayed on a Smith Chart. **Focus Microwaves, Inc., 1603 St. Regis, Dollard-des-Ormeaux, Quebec H9B 3H7, Canada; (514) 683-4554, FAX: (514) 684-8581, e-mail: christos@focus-microwaves.com, Internet: www.focus-microwaves.com.**

REFERENCE

1. M. Demmler, B. Hughes, and A. Cognata, "A 0.5-50GHz on-wafer intermodulation load pull and power measurement system", Proceedings of the Microwave Theory & Techniques Symposium, 1995, Orlando, FL.

High Linearity Direct Conversion Receivers



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	LT5515	LT5516
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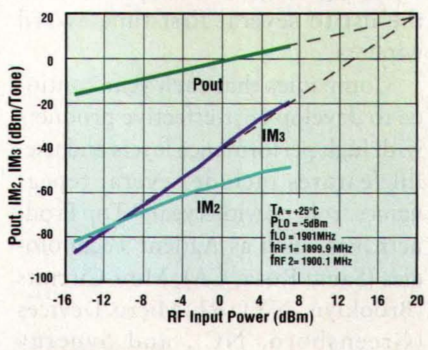
IIP2	51dBm	52 dBm
Noise Figure	16.8 dB	12.8 dB
Conversion Gain	-0.7 dB	4.3 dB
LO-RF Leakage	-46 dBm	-65 dBm
LO Drive Level	-5 dBm	
Supply Voltage	5V	
Package	4mm x 4mm QFN	

I/Q Output Power, IM2, IM3 vs RF Input Power



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PRODUCT *technology*

Top Products *of 2003*

This diversified collection of hardware, software, and test equipment represents some of last year's best creations using existing and cutting-edge technologies.

Difficult times notwithstanding, 2003 marked the arrival of many innovative products in hardware, software, and test equipment. Selected by the editors of *Microwaves & RF* for their technical innovation and practical benefits to designers (see table), the Top Products of 2003 represent the many efforts of gifted design engineers, test engineers, technicians, and manufacturing professionals during a year when few companies took a single product sale for granted. As a sign of the industry's health and positive long-term prospects, however, the list also shows a diversity of companies, from repeating "regular" contributors to the list to several first-time award winners.

JACK BROWNE
Publisher/Editor

Companies that each year continue to develop cost-effective products with high performance levels and useful features include several repeat names from previous years' Top Products lists, such as Agilent Technologies (Santa Rosa, CA), Mini-Circuits (Brooklyn, NY), RF Micro Devices (Greensboro, NC), and Synergy Microwave (Paterson, NJ). Although the range of their products extends from the smallest of 90-deg. power splitters (from Mini-Circuits) to sophisticated (and not inexpensive) computer-aided-engineering (CAE) software suites (from Agilent Technologies), the criteria for selecting these products is the same: these are new products that allow engineers to approach their jobs and create designs more easily and efficiently, to save engineering/product cost and size, and to achieve new levels of performance.

The RF Design Environment (RFDE) software tools from Agilent Technologies, for example, provide system-level verification of RF circuit designs and physical-level modeling of components within a circuit layout. The design and verification functionality, developed in conjunction with Cadence Design Systems (San Jose, CA), is meant to improve the efficiency of the integrated-circuit (IC) design process.

Some of the company names on this year's list are relative newcomers to the top-product notoriety, even though these are well-established firms with solid track records. The PN9002 radar-stability test system from Aeroflex, Inc. (Plainview, NY), for example, offers a different look at test equipment,

designed around plug-in modules fitting into two VXI racks. With a standard (military) frequency range of 2 to 18 GHz (and optional range of 0.4 to 18 GHz), the test system can check the amplitude and phase stability of radar pulses at pulse widths as narrow as 200 ns. The $160,000 system can replace custom measurement solutions costing more than $1 million. Additional outstanding measurement solutions represented on the list include a new line of Infinity Probes from Cascade Microtech (Beaverton, OR) for on-wafer testing of the next-generation of high-frequency silicon-based devices, including those fabricated on silicon-germanium (SiGe) substrates, and the wideband CCMT-6510 65-GHz coaxial impedance tuner from Focus Microwaves (Dollard des Ormeaux, Quebec, Canada), a marvel of mechanical and electrical engineering. For engineers working with electromagnetic-compatibility (EMC) testing, the unique Radiant Arrow 26 "bent-element" antenna design from Amplifier Research (Souderton, PA) provides frequency coverage down to 26 MHz (and to 5 GHz at the high end) without occupying the space of conventional log-periodic antennas.

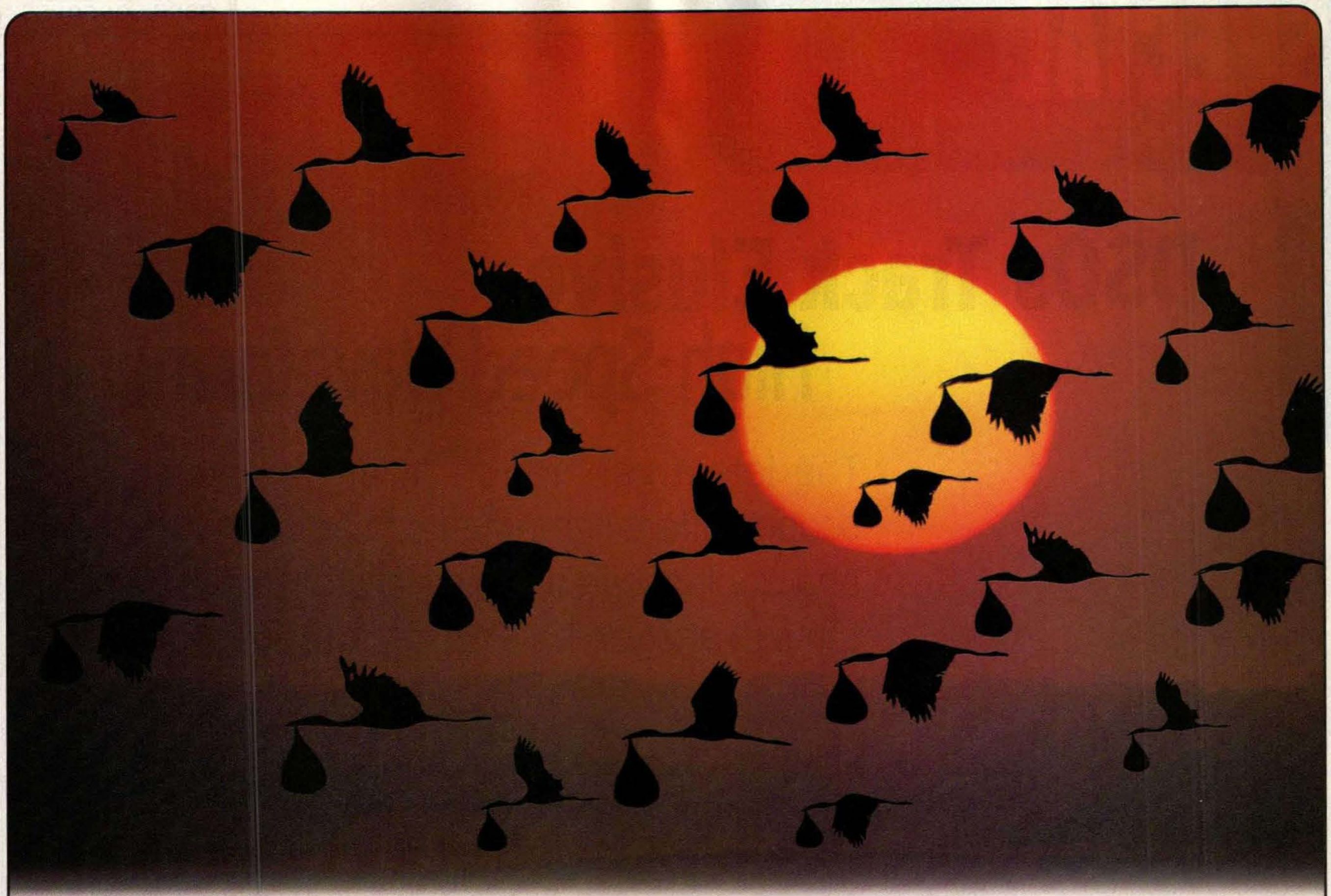
The ADV-3000S DDS source from newcomer Advanced Radio Corp. (Reston, VA) offers microsecond switching speed from 20 MHz to 3 GHz (and can be extended to 18 GHz) with unheard-of spurious performance of better than –90 dBc for a DDS. Ideal as a local oscillator (LO) for signal-intelligence (SIGINT), electronic-warfare (EW), and radar systems, the source boasts low phase noise, 1-Hz frequency resolution, and the capability to generate a wide range of modulation formats, including quadrature amplitude modulation (QAM) and quadrature phase-shift-keying (QPSK) modulation. Also in the area of signal generation, the CRO line of ceramic resonator oscillators from Synergy Microwave Corp. (Paterson, NJ) offers new levels of performance from an existing technology, while the MEMS-based oscillators from Discera (Campbell, CA) stand as a radical departure from existing oscillator technologies and perhaps an eventually replacement for the long-trusted quartz crystal oscillator. In addition, SiGe direct quadrature modulators from Hittite Microwave Corp. (Chelmsford, MA) offer enough bandwidth (as much as 3 GHz) for the most exotic modulation formats, while one of the first source products from startup Centellax (Santa Rosa, CA) is the industry's highest-frequency fractional-N synthesizer, employing a merchant SiGe process to generate clean signals to 30 GHz.

As a sign of the growing WLAN market, several chip sets were included on the 2003 list, including RF and baseband devices from RF Micro Devices and integrated solutions for both 2.4- and 5-GHz WLAN systems from IceFyre Semiconductor (Kanata, Ontario, Canada). IceFyre's SureFyre chips (for 5-GHz 802.11a systems) and TwinFyre ICs (for 2.4- and 5-GHz 802.11a/b/g systems) include a unique Class F switch-mode GaAs PHEMT amplifier capable of +21 dBm output power at 5 GHz with better than 35-percent power-added efficiency.

Another innovative amplifier technology represented on the list is the grid-array millimeter-wave power amplifiers from Wavestream Corp. (West Covina, CA). Also based on a GaAs PHEMT process, the amplifiers, which use spatial-combining techniques rather than passive hybrids, have achieved +36 dBm output power at 1-dB compression at 38 GHz from a single chip. Finally, the barium-strontium-titanate (BST) materials from Agile Materials & Technologies (Goleta, CA) may represent one of the most groundbreaking developments during 2003. Given that these materials can change dielectric constant with applied voltage, these dielectric materials should give rise to a new generation of voltage-tunable devices, such as delay lines and phase shifters, appearing on future Top Product lists. MRF

Top Products of 2003 at a glance (listed alphabetically)

1. *Advanced Radio Corp.'s* ADV-3000S 20-MHz-to-3-GHz direct digital synthesizer (April, p. 94)
2. *Aeroflex's* PN9002 VXI-based radar-stability test system (May, p. 120)
3. *Agile Materials & Technologies'* barium-strontium-titanate (BST) tunable-dielectric-constant materials (November, p. 106)
4. *Agilent Technologies'* RF Design Environment (RFDE) CAE design tools (November Cover, p. 86)
5. *Amplifier Research's* Radiant Arrow 26 26-MHz-to-5-GHz EMC antenna (June, p. 102) (amplifiers.com)
6. *Cascade Microtech's* 110-GHz Infinity wafer probes (April Cover, p. 84)
7. *Centellax's* SiGe-based 30-GHz fractional-N synthesizer operates to 30 GHz (December Cover, p. 88)
8. *Discera's* MRO-100 microelectromechanical-systems (MEMS) oscillator (August Cover, p. 84)
9. *Focus Microwaves'* CCMT-6510 65-GHz harmonic load-pull tuner (January Cover, p. 97)
10. *Hittite Microwave Corp.'s* HMC495LP3 and HMC496LP3 SiGe direct quadrature modulators (October Cover, p. 78)
11. *IceFyre Semiconductor's* SureFyre and IceFyre 802.11a/b/g WLAN chip sets (October, p. 100)
12. *Mini-Circuits'* QCC-20 and QCC-22 LTCC-based 90-deg. splitters (June Cover, p. 90)
13. *RF Micro Devices'* RFCS5420 802.11a/b/g WLAN chip set (May Cover, p. 108)
14. *Synergy Microwave Corp.'s* CRO family of ceramic resonator oscillators (September Cover, p. 100)
15. *Wavestream Corp.'s* grid-array millimeter-wave power amplifiers (July, p. 94)



DSOs Track Elusive High-Speed Waveforms

The latest generation of digital oscilloscopes offers lightning-fast sampling rates and sophisticated timing and display systems to analyze hard-to-capture electronic events.

JACK BROWNE
Publisher/Editor

Oscilloscopes are not often considered the main measurement tools of the RF/microwave engineer. However, with the increasing use of complex modulation formats in commercial communications, and "exotic" signal formats in military systems, the digital storage oscilloscope (DSO) is now a vital test instrument throughout the high-frequency industry. Understanding some of the key performance specifications can help simplify the task of finding the right DSO for the job.

Digital oscilloscopes have all but replaced analog models where performance, flexibility, and functionality are critical, although older analog units are still available from some sources, such as on-line auction firms such as Ebay.com and rental/leasing companies such as Electro Rent Corp. (www.electrorent.com). Experienced RF engineers may recall setting up a Polaroid instant camera on a hood in front of an analog oscilloscope, trying to freeze the image of a fast-moving waveform. Since a DSO captures a signal by means of sampling through an analog-to-digital converter (ADC), any captured signal information exists in digital form and can be saved and recalled at a user's convenience. Many modern DSOs include analog-like display capabilities, such as a persistence mode, to simplify operation for users familiar with older analog oscilloscopes.

DSOs can be compared in terms of various key performance parameters, including those related to the instrument's display, time base, and acquisition/triggering capabilities. Display specifications usually include the nominal analog input bandwidth, the typical rise time of a measurement channel, the number of channels, the sensitivity (usually in millivolts per display division), the maximum input voltage, the vertical resolution (in bits), DC gain accuracy (as a percentage of the full-scale reading), and offset range and accuracy. The time base or horizontal system, which is tied to the quality of the internal reference source (usually a stabilized crystal oscillator), is usually characterized in terms of the time/division range (usually from ps/division to s/division), the clock accuracy (in PPM), and the jitter noise floor. The acquisition system, which handles how input signals are sampled and stored in memory, is usually characterized in terms of the single-shot sample rate per channel (the capability of capturing a single random event), the interleaved sample rate (in which the sampling power of ADCs from two channels are combined to effectively double the single-shot rate), and the random interleaved sample rate (which provides high sample rates for

1. The four-channel WaveRunner 6100 DSO sports 1-GHz analog bandwidth and single-shot sampling rate of 5 GSamples/s, which can be interleaved to 10 GSamples/s for two-channel measurements. (Photograph courtesy of LeCroy Corp., Chestnut Ridge, NY.)

capturing more predictable, repetitive events).

Although oscilloscopes have traditionally been classified as either high-performance or general-purpose units, the gap between the two types continues to narrow. The Infiniium series of DSOs from Agilent Technologies (www.agilent.com) includes not only general-purpose (with bandwidths from 600 to 1000 MHz) and high-performance models (bandwidths from 2250 to 6000 MHz), but also portable instruments (from 60 to 500 MHz), models for mixed-signal use (from 60 to 1000 MHz), and even instruments that combine a digital communications analyzer (DCA), oscilloscope, and time-domain reflectometer within a single mainframe (with bandwidths from 3 to 80 GHz). The Infiniium instruments typically provide four measurement channels with 12-b vertical resolution and less than 1-ps root-mean-square (RMS) time base jitter. The DCA models, which provide specialized communications testing with data presented on eye diagrams and constellation displays, can be configured with a wide range of electrical and electro-optical modules to meet the needs of wired, wireless, and optical communications systems.

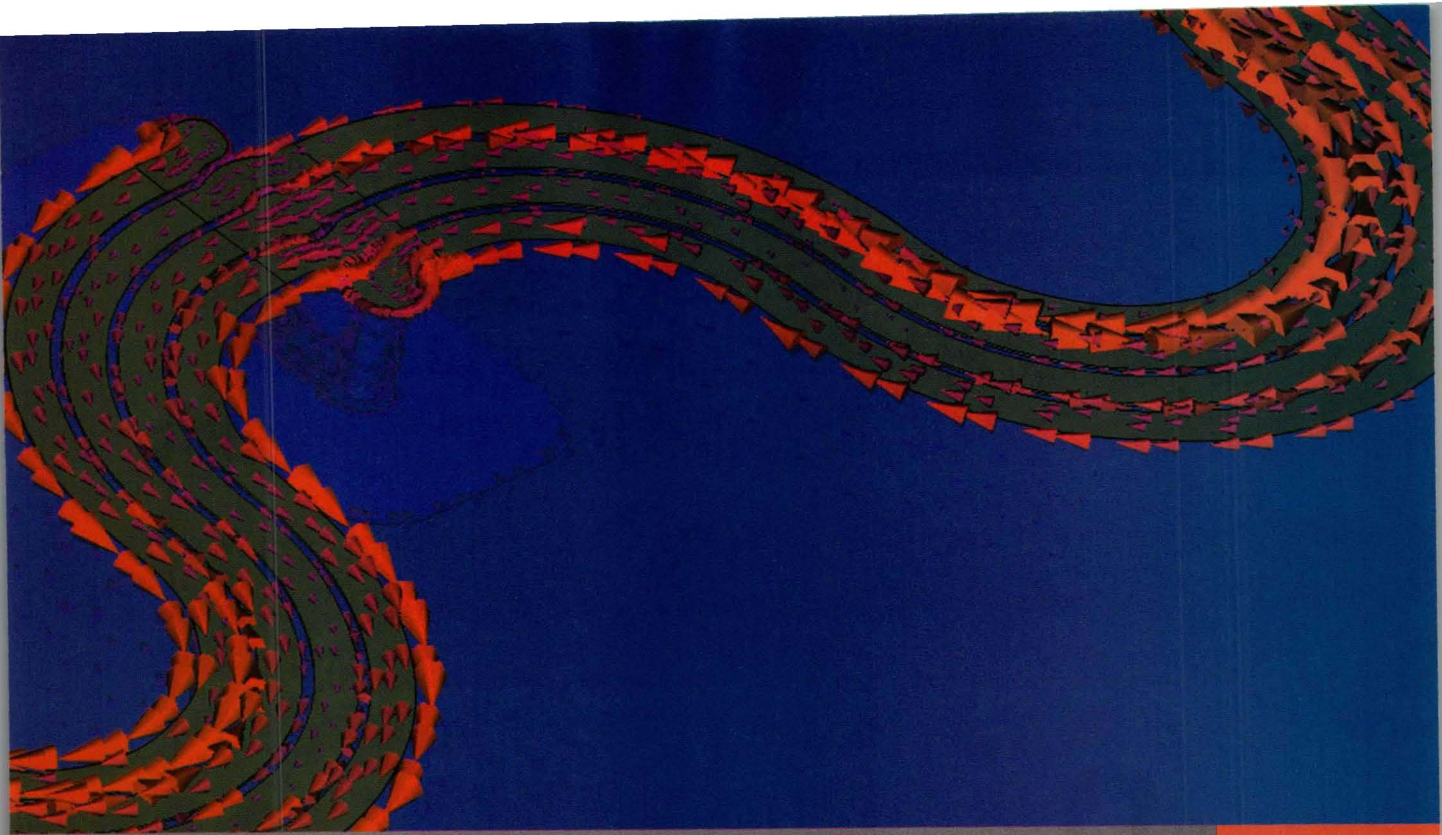
The WaveRunner 6000 series DSOs from LeCroy Corp. (Chestnut Ridge, NY) are sold as general-purpose scopes, but provide man of the features and performance levels associated with high-performance instruments. For example, all models include 8.4-in. color touch-sensitive screens, along with at least 1 million points of memory per channel, and 5 PPM time base stability; options provide 2, 4, 8, or 12 Mpoints of memory per channel. Models include the 500-MHz WaveRunner 6050, the 1-GHz WaveRunner 6100 **(Fig. 1)**, and the 2-GHz WaveRunner 6200. These typically four-channel oscilloscopes feature a 5-GSamples/s ADC on each channel. The ADCs from two channels can be interleaved to create a single 10 GSamples/s channel (resulting in a total of two measurement channels instead of four). The WaveRunner instruments include an analog persistence feature to create an analog-like display that can ease the analysis of some measurement data (such as video signals). For more high-performance applications, the company also offers the WaveMaster 8000A Series with single-shot sampling rates to 20 GSamples/s and repetitive interleaved sampling to 200 GSamples/s at bandwidths to 6 GHz and the 1-to-3-GHz midrange WavePro Series oscilloscopes.

2. The TDS7704B Digital Phosphor Oscilloscope (DPO) features a 7-GHz analog bandwidth and 20 GSamples/s single-shot sampling rate. (Photograph courtesy of Tektronix, Inc., Beaverton, OR.)

Another company that is migrating high performance levels to more "general-purpose" applications is Tektronix (www.tektronix.com), with its advanced Digital Phosphor Oscilloscope (DPO) technology in which capture rates exceeding 400,000 waveforms per second are used to instantaneously collect and analyze data about high-speed signals and provide meaningful two- and three-dimensional displays of test results. The company's TDS7000B series of DPOs include a triggering system based on silicon-germanium (SiGe) components and trigger jitter as low as 1 s RMS. The flagship model TDS7704B DPO features a 7-GHz input analog bandwidth and 43-ps rise time along with a 20 GSamples/s single-shot sampling rate well suited for analysis of 4.25-Gb/s serial data **(Fig. 2)**. As with Agilent Technologies, Tektronix also offers a comprehensive line of communications signal analyzers (CSAs) with signal processing and display capabilities designed to provide meaningful displays of captured communications signals, and the new RSA Series real-time spectrum analyzers which are designed to capture and digitize signal bandwidths as wide as 10 MHz at frequencies to 8 GHz. The RSA instruments also offer a DSO-like time-domain mode based on the use of Fast Fourier Transform (FFT) capability.

While the instruments mentioned so far represent traditional, stand-alone units, several oscilloscope suppliers leverage personal-computer (PC) technology for their products, including Acqiris Technology (www.acqiris.com), Gage Applied Sciences (www.gage-applied.com), and Pico Technology (www.picotech.com). For example, the model DP240 PCI digitizer card from Acqiris provides the characteristics of a DSO for two measurement channels, a 1-GHz bandwidth, and a 2 GSamples/s sampling rate. The company's DC271 CompactPCI digitizer offers four channels, a 1-GHz analog bandwidth, sampling rates from 1 to 4 GSamples/s, and as much as 32 Mpoints acquisition memory. The CompuScope 82G PCI digitizer card from Gage is now available with optional 1-GHz bandwidth and 1 GSamples/s sampling rate simultaneously on two channels. The card can be equipped with as much as 16 MB of acquisition memory.

Several suppliers offer analog and digital oscilloscopes designed for portability, generally with bandwidths of 100 MHz or less. These include the three-channel, 100-MHz, 100-MSamples/s model LS8106A oscilloscope from Leader Instruments (www.leaderusa.com) and the 100-MHz D7510 Series DSOs from HC Protek (www.hcprotek.com). Perhaps the most compact and portable digital oscilloscopes currently available are the ScopeMeter instruments from Fluke (www.fluke.com), which also include digital-multimeter (DMM) capability. For example, the battery-powered, handheld model 199C features a 200-MHz input bandwidth, 2.5 GSamples/s sampling rate, and 1200 points of acquisition memory, with captured information shown on a color screen with digital and variable-persistence display modes.



Synthesizers Switch To 40 GHz In Nanoseconds

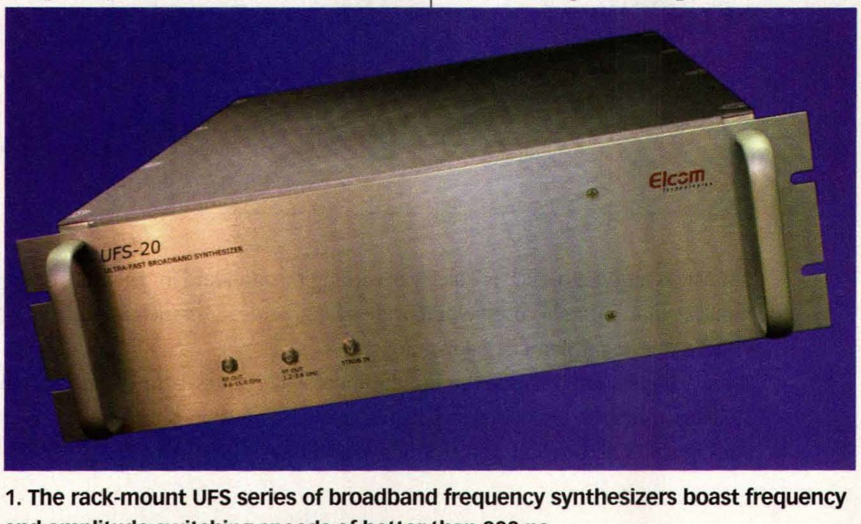
These direct-analog sources maintain good spectral purity and high frequency resolution over broad bandwidths for a variety of demanding military applications.

JACK BROWNE
Publisher/Editor

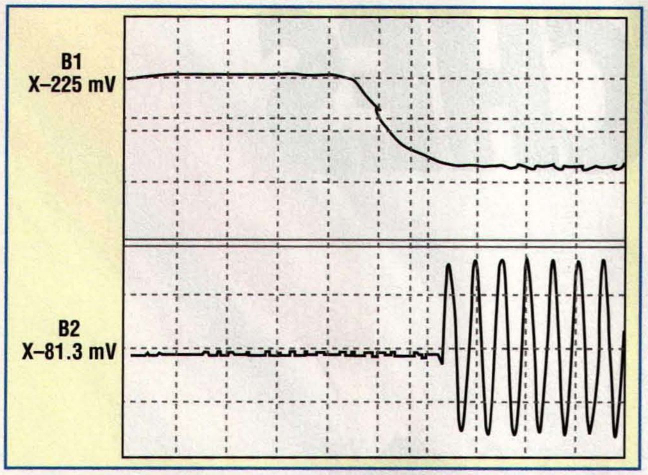
Nanosecond synthesizer switching speeds are critical in many military applications, including radar, electronic-warfare (EW), and surveillance systems. The UFS Series of frequency synthesizers from Elcom Technologies (Rockleigh, NJ) bring this kind of frequency and amplitude switching speed over the wide bandwidths required by military integrators, from 0.01 to 40.00 GHz. The rugged, rack-mount synthesizers are based on direct-analog frequency-synthesis techniques and a basic module design that supports rapid customization for specific system requirements.

The UFS-18 is an example of the UFS Series, with a frequency range of 300 MHz to 18 GHz. The switching speed—whether to settle to a new frequency or within 2 dB of a new amplitude setting—is a mere 200 ns. Frequency resolution can be specified as fine as 1 Hz, depending on requirement, and typical output power is +10 dBm with ±2 dB output-power flatness across the full frequency range and an operating temperature range of 0 to +50°C.

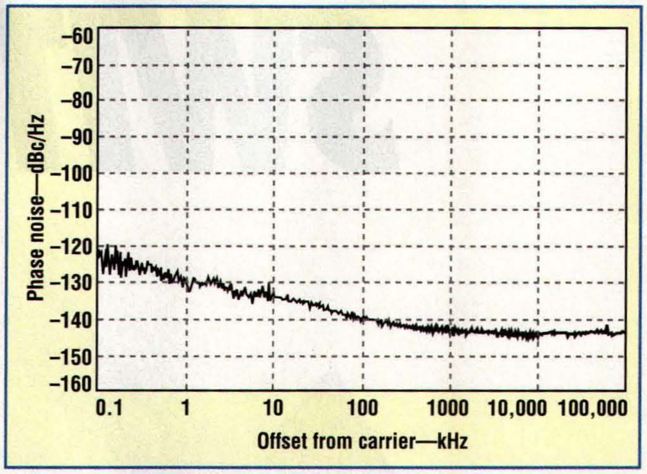
Of course, switching speed without good spectral purity would introduce false readings in many military systems, so the UFS synthesizers **(Fig. 1)** are designed to operate with low



1. The rack-mount UFS series of broadband frequency synthesizers boast frequency and amplitude switching speeds of better than 200 ns.



2. Frequency switching speed was measured at 105 ns for a transition from 9610 to 9643 MHz. The display scale is 50 ns/division.



3. Residual phase noise for the UFS-15 offset from a 3-GHz carrier was measured with an E5500 phase-noise measurement system from Agilent Technologies (Santa Rosa, CA).

phase noise, harmonics, and spurious levels. The UFS-18 exhibits low harmonic levels of –50 dBc, with maximum spurious levels of –70 dBc from 0.3 to 18 GHz. The single-sideband (SSB) phase noise is –60 dBc/Hz offset 10 Hz from a 10-GHz carrier, dropping to –90 dBc/Hz offset 100 Hz from the same carrier, –126 dBc/Hz offset 10 kHz from the same carrier, and reaching –140 dBc/Hz offset 10 MHz from a 10-GHz carrier. SSB phase noise in low-noise version is dropping to –147 dBc/Hz offset 10 MHz from the same 10-GHz carrier. A low-noise option further reduces the SSB phase noise floor to a pristine –147 dBc.

The synthesizer is equipped with an internal oven-controlled crystal oscillator (OCXO), which delivers frequency accuracy of 1 PPM. The synthesizer can also phase lock to an external 5- or 10-MHz reference source, assuming the frequency accuracy of that source in the process. The UFS-18 weighs 60 lbs. and measures 5.22 × 16.75 × 22 in. (13.3 × 42.5 × 55.9 cm). It can be programmed remotely by means of 44-b parallel binary-coded-decimal (BCD) commands (the company provides easy-to-use, Windows-based software for ease of programming); GPIB programmability is available as an option on this model as well as other frequency derivatives.

The UFS Series synthesizers are well suited for any applications requiring fast measurement throughput or high-speed frequency/amplitude switching, including radar cross-section measurements, antenna testing, EW LO generation, and wideband surveillance monitoring. They are designed for operating temperatures from 0 to +50°C and can be supplied with a variety of optional modulation formats, including wideband amplitude modulation (AM), wideband frequency modulation (FM), phase modulation, and (10-ns-risetime) pulse modulation. The wideband FM supports modulation bandwidths of DC to 6 MHz in DC mode and 3 kHz to 6 MHz in AC mode. FM deviation can be controlled over a range of ±100 MHz peak-to-peak at rates of 250 MHz/μm. Digital modulation, which can be used to generate digital chirp output signals, is available with an instantaneous bandwidth of 40 MHz. the fast-risetime pulse modulation is available with on/off ratios of 80 dB to 18 GHz. Along with the wide array of modulation possibilities, the UFS frequency synthesizers can be packaged according to a customer's requirements, including in VXi configuration, with a front-panel keypad and display for data entry, and with front-panel SMA RF output connectors.

***UFS Series** synthesizers are well suited for any applications requiring fast measurement throughput or high-speed frequency/amplitude switching.*

In addition to the UFS series, the company also offers the WMFS series of module-format synthesizers for military applications from 1 to 23 GHz (in 6-GHz bands), the MFS Series of sources for satellite-communications applications, the DFS series of 1-to-23-GHz (in bands) synthesizer modules for digital radios, and the low-power broadband IBS series of VME synthesizers for commercial and military use from 0.05 to 18.0 GHz. The WMFS series synthesizers, for example, feature output power to +14 dBm with 1-Hz frequency resolution, –70 dBc typical spurious performance, and –95 dBc/Hz phase noise offset 100 kHz from a 10-GHz carrier. **Elcom Technologies, 11 Volvo Dr., Rockleigh, NJ 07647; (201) 767-8030, FAX: (201) 767-6266, e-mail: info@elcom-tech.com, Internet: www.elcom-tech.com.**

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MCA1-85	7	2800-8500	5.6	38	8.95
MCA1-12G	7	3800-12000	6.2	38	10.95
MCA1-24LH	10	300-2400	6.5	40	6.45
MCA1-42LH	10	1000-4200	6.0	38	7.45
MCA1-60LH	10	1700-6000	6.3	30	8.45
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earlier.

Following similar arguments as in Case 1, the ratio of the efficiencies of OCFF and FF can be shown as:

$$\frac{\eta_{OCFF}}{\eta_{FF}} = \frac{1}{l_m} \times \frac{\dfrac{1}{\eta 1} + \dfrac{1}{\eta 2} \times \dfrac{\delta}{c_2}}{\dfrac{1}{\eta 1} + \dfrac{1}{\eta 2} \times \dfrac{\delta \times o_f}{c_2 \times g_{A1}}} \quad (10b)$$

For practical applications, $g_{A1} > o_f$ and as in case 1, the efficiency of an OCFF system is always larger than that of an FF system.

From the above analysis, the improvement in efficiency offered by the OCFF approach compared to the traditional FF approach is evident, where the effects of eliminating main line loss and using a smaller error amplifier (A2) were demonstrated. The OCFF can use a higher-loss and lower-cost delay line than the FF approach, such as a lumped-element delay line. The lower-power dissipation of A2 translates into lower cost and smaller size for the overall system.

Experiments performed with Class A amplifiers demonstrated the OCFF concept. The output third-order intercept of main amplifier A1 and driver amplifier were +35.5 and +30.5 dBm, respectively. The main path delay line exhibited loss of about 2.5 dB at 800 MHz and delay of about 10 ns. Results for both approaches are shown in **Figs. 5(a) and 5(b)**. Without linearization, amplifier A1 produces third-order intermodulation distortion levels of –40 dBc in the 700-MHz band and –42 dBc in the 800-MHz band for output power of +15 dBm per tone. Using the OCFF linearizer, suppression levels for third-order intermodulation distortion were more than 18 and 15 dB in the 700- and 800-MHz bands, respectively.

The OCFF technique provide comparable linearization performance to the FF approach. The technique is especially attractive since it eliminates any loss at the output of the power device from delay line and coupler as in a FF scheme. It also improves the overall efficiency of the system compared to a FF scheme. The technique can be applied where the output is of a different nature from the input, as in electrical to optical conversion, or a transmitter stage involving frequency upconversion with linearization is performed at intermediate frequencies (IFs). The OCFF approach allows the linearization of direct modulated lasers where the distortion depends on the link distance due to the interaction of the laser chirp with fiber dispersion. There is no limitation on the bandwidth of the OCFF approach, and the technique can be implemented where both even- and odd-order distortion components must be controlled. MRF

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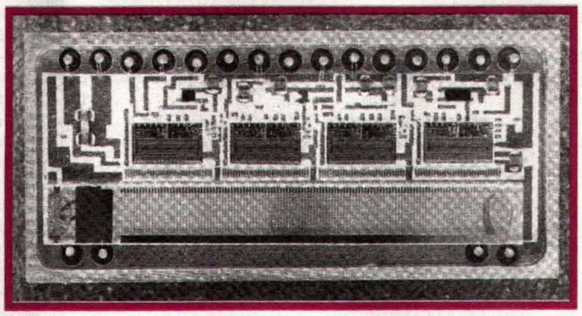

ALMOST 15 YEARS AGO, designers from Hazeltine Corp. (Greenlawn, NY) developed a tiny surface-acoustic-wave (SAW) filter with integral control application-specific integrated circuits (ASICs) capable of 128 different minimum-shift-keying (MSK) phase states at a center frequency of 80 MHz.

next month

Microwaves & RF February Editorial Preview

Issue Theme: Semiconductors

News

February wraps up Microwaves & RF's three-issue preview coverage of the upcoming Wireless Systems Design Conference & Expo scheduled for March 8-10, 2004 in the San Diego Convention Center (San Diego, CA). This final preview article will highlight many of the technical presentations to be offered as part of dual educational platforms on commercial wireless technologies as well as military electronic systems design strategies. The Special Report will also feature a sneak preview of some key new products to be unveiled at this 12th annual wireless event. February will also feature a summary of the 62nd Automatic RF Techniques Group (ARFTG) meeting on differential measurement strategies.

Design Features

Several articles inspired by wireless-based design issues support February's Semiconductor theme. In one, an author from a leading CAE supplier will explain how to improve the phase noise of a VCO. In another, an author for a major device manufacturer addresses the concerns in assembling hardware for remote-keyless-entry (RKE) systems. In addition, an author from a top analog device supplier will explore design options for multicarrier cellular transmitters Also, the issue will include the second installment of a four-part article series on designing FIR digital filters and techniques for characterizing reed relays through 10 GHz.

Product Technology

February features a generous selection of new product developments, starting with a line of high-performance 12- and 15-b ADCs and how their precision and speed support modern digital-receiver designs. Additional Product Features will introduce the industry's first complete monolithic dual-band GSM handset power amplifier based on standard silicon CMOS, a line of low-noise, general-purpose amplifier gain blocks well suited for applications from 0.1 to 6.0 GHz, a versatile transceiver IC that works for the three modes and two frequency bands of 802.11a/b/g WLAN systems, and a family of lowpass filters that cover 0.6 to 3.0 GHz.